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Like other human institutions, engineering is in a process of continuous evolution. One fundamentally important broadening of the scope, opportunities, and obligations of the engineer is forcefully presented in the following editorial by the Secretary of the Institute (who, it may be mentioned, is himself an outstanding instance of the successful blending and performance in an individual of the functions of the technologist and the industrialist).—*The Editor.*

Society's Hopes for the Engineer

HARADEN PRATT

Never before have business and industry been so alerted to the importance of technical research and development. This consciousness now even extends into the body politic generally. The daily papers and most popular magazines found in every home have taken up the theme and dwell upon the wondrous achievements recently made in weapons, machines, and basic science applications. We read about the coming industrial revolution and the effects that it will have on the methods of living with its new materials, advanced manufacturing processes, easy communications, and spectacular means of transport.

All of this rehearsing and speculation is conditioning the people to the realization that we must advance technically or be left behind in the future world. They are being taught that a nation not adequately sponsoring basic research and technical development will become a back number and eventually be at the mercy of the others. But to be successful such a nation must maintain an adequate educational system whereby it can produce the trained people required, a free and encouraging atmosphere in which they can work, a universal exchange of technical information, suitable incentives without which best effort cannot be achieved, and a broadened point of view, both by and towards the engineer.

An industry, just as a nation, can no longer blunder ahead successfully leaving the technological aspects to chance. It must plan carefully. It cannot proceed far with such plans without expert technical guidance. Forward-looking business administrators will be driven to seek out technically informed persons to become active if not leading elements of their directing organizations.

The new awareness of the indispensability of the technical man featuring this postwar era and the now generally prevalent public enthusiasm of expectation create a serious challenge to our engineering profession because it will be the engineer who must become acquainted with the reservoir of scientific knowledge, appraise it in the light of the needs of the life of the people, and then, by judicious interpretation, develop the useful applications. The engineer must better fit himself to meet this obligation and it is to this new opportunity and its problems that I wish to call particular attention.

The experience of the technically trained man gained in the molding together of the tangible and intangible provides him with a unique point of view which is urgently needed to complement the thinking of legislators and the pioneers of enterprises. Too few engineers realize this broadened role into which they must expand. They must themselves take advantage of the occasion to reorient their objectives and recognize that the highest attainment in their profession is to become masters of the fundamentals and facts of science and at the same time to discern the cry of the people and minister to their wants. They must themselves take the lead in enlarging the engineer's scope and educate themselves to fit it, thereby not only contributing to their own success but ensuring the better fulfillment of their proper function in the society of today and tomorrow.

Transmission of Television Sound on the Picture Carrier*

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Summary—Several pulse methods for the transmission of television sound on the picture carrier during the line-blanking intervals are analyzed from the points of view of signal-to-noise ratio, audio fidelity, and transmitter and receiver design.

The advantages of duplex transmission are: (1) elimination of a separate sound transmitter; (2) elimination of the ambiguity and difficulty which may occur when a standard frequency-modulated sound signal is tuned in; (3) freedom of the audio output from the type of distortion which occurs in frequency-modulated receivers as a consequence of excessive drift of the frequency of the local oscillator; and (4) improvement of the phase characteristic of the picture intermediate-frequency amplifier resulting from elimination of trap circuits.

The highest audio-modulation frequency in duplex systems must not exceed one half of the line-scanning frequency. This is a disadvantage under the present television standards which specify a line frequency of 15,750 cycles per second.

With the exception of pulsed frequency modulation, the signal-to-noise ratios of sound in duplex systems are not so great as the ratio offered by the transmission of a standard frequency-modulated carrier. The comparison is subject to the condition that the amplitude of the frequency-modulated carrier is 0.7 of the peak amplitude of the duplex carrier. The signal-to-noise ratio of a pulsed frequency-modulated signal may equal the ratio of a standard frequency-modulated signal up to a critical distance from the transmitter, but is less at greater distance.

Introduction

FROM TIME to time, proposals have been made that the sound accompaniment for television may be transmitted by a modulation of the picture carrier during the line-blanking intervals when no picture detail is transmitted. Improved reception of picture and sound, decreased investment in receivers and transmitters, and greater channel width for the picture signal are mentioned as possibilities of a "duplex" transmission of picture and sound. The purpose of this paper is to assist engineers in their evaluation of duplex transmission as a practicable system by offering an analysis of several methods of duplexing.

A review of the method of sound transmission and reception in use is helpful as a setting for the discussion. In the present arrangement, sound is transmitted by a frequency-modulated sound transmitter operating on a carrier frequency which is 4.5 megacycles above the picture carrier. At the position of the sound carrier, the recommended standards[1] state that the field strength of the picture sidebands shall not exceed 0.0005 of the picture carrier. There is essentially nothing at the transmitting point to distinguish a television sound transmitter from a conventional frequency-modulation transmitter designed for operation in the frequency-modulation band. The sound receiver is likewise conventional and may share only the same heterodyne oscillator with the picture receiver.[2]

The picture transmitter is amplitude-modulated and radiates a wave form illustrated by Fig. 1. Picture content is transmitted during about 85 per cent of the total "time on the air." No picture detail is transmitted during the blanking intervals of the scanning tubes in the transmitter and receiver. Such "idle" intervals amount to about 15 per cent of the total time.

Proposals[3,4] for duplexing have been directed, therefore, toward the utilization of some part of the blanking interval for sound transmission. Thus the television transmitter would be converted into a picture-sound duplex transmitter which radiates picture intelligence during 85 per cent of the time and sound intelligence during some part of the remaining 15 per cent, using only one antenna and one radio-frequency power amplifier. There would be a synchronized electronic switch in the receiver for the opening of the sound channel to the video signal sometime during the blanking interval.

Factors which are involved in a comparison of duplex methods and the present method of continuous transmission of sound are the following: (1) audio fidelity; (2) signal-to-noise ratio; (3) amount of interaction between video and audio signals; (4) permanency of receiver alignment; (5) picture quality; (6) ease of receiver tuning; (7) cost of receiver; (8) cost of transmitter.

Certain advantages and disadvantages of a duplex system may be predicted in advance of a theoretical and experimental analysis. First, there is no separate sound transmitter and antenna. This economic advantage can not be accorded much weight unless there is a resultant economy in the television receiver because the ratio of receivers to transmitters is so great that the economics

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‡ Columbia Broadcasting System, New York, N. Y. Work covered in this paper was done while the author was a member of RCA Laboratories, Purdue University, Lafayette, Ind.

[1] Final report of the Radio Technical Planning Board.

[2] In some designs, the sound and picture signals are amplified at intermediate frequency in one or more common stages before branching off into separate intermediate-frequency amplifiers occurs.

[3] H. E. Kallman, "Audio and video on a single carrier," *Electronics*, vol. 14, pp. 39–42; May, 1941.

[4] Numerous patents including: U. S. patents, No. 1,655,543, R. A. Heising; No. 1,887,237, J. L. Finch; No. 2,061,734, R. D. Kell; No. 2,083,245, H. Shore and J. N. Whittaker; No. 2,086,918, D. G. C. Luck; No. 2,089,639, A. V. Bedford; No. 2,227,108, H. A. Rosenstein; No. 2,257,562, H. Branson; No. 20,153, E. F. W. Alexanderson (reissue).

of the receiver is the controlling factor.

A significant advantage of a duplex receiver is the freedom of the audio output from the type of distortion which occurs in frequency-modulation receivers as a consequence of excessive drift of the frequency of the local oscillator. Audio modulation is conveyed in a duplex system by the envelope of a radio-frequency carrier and is relatively unaffected by instability of the local oscillator.

Sound-rejector circuits in the picture intermediate-frequency amplifier can be removed with a resulting reduction in phase distortion and some improvement in picture quality.

The ambiguity involved in tuning-in a conventional frequency-modulation signal is removed. In the present system a choice must be made of tuning for minimum interference in the picture or minimum noise in the sound in receivers which are somewhat misaligned.

There is also available a small extension of the video band into the space now assigned as a guard band for the sound; this amounts to about one quarter of a megacycle.

It may be anticipated that the signal-to-noise ratio with duplex sound would be unfavorable as a consequence of the reduced time for transmission of sound. This is a serious obstacle in some duplex systems. A further limitation is the imposition of a maximum audio frequency that may be transmitted without the introduction of spurious frequencies into the audio spectrum. This restriction has been noted before in pulse transmission of sound. Finally, there is the complication of synchronizing the electronic sound selector in the receiver with the line-scanning frequency.

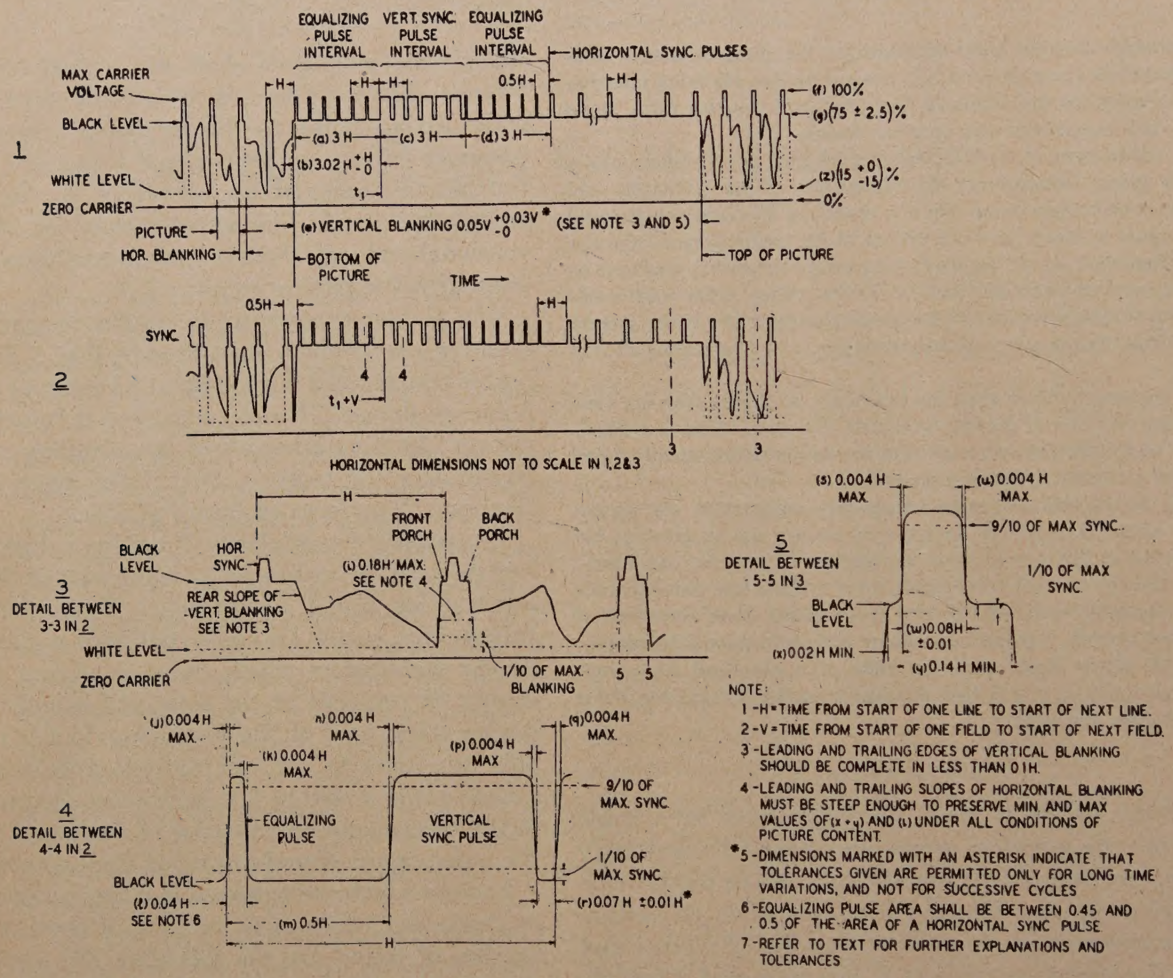


Fig. 1—Television synchronizing wave form.

Duplex Methods

1. *Amplitude-Modulated Pulses*

One of the most obvious duplex systems is the modulation of the amplitude of a rectangular pulse wave in accordance with the audio signal and the insertion of the modulated pulses in the line-blanking interval of the video signal. Fig. 2 illustrates the successive steps at the transmitter. The pulse wave form shown in (A) of Fig. 2 is amplitude modulated by the audio signal $M(t)$, as

illustrated in (B). In (C) the modulated pulses have been inserted in the part of the blanking interval following the synchronizing pulse. Such a composite wave would be applied as modulation of the picture carrier.

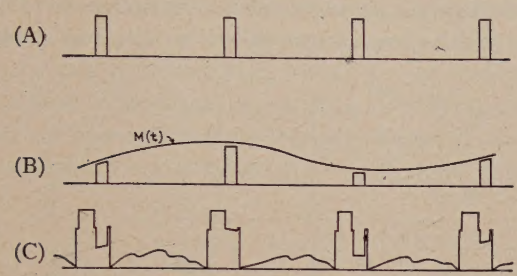


Fig. 2—Amplitude-modulated duplex wave forms.

(A) pulse carrier
(B) audio wave form $M(t)$ and amplitude-modulated pulse carrier
(C) amplitude-modulated pulses combined with television wave form

If it is granted that the synchronized electronic switch in the receiver is able to select the pulses from blanking so that the pulse wave in Fig. 2(B) is recovered, the audio fidelity of the duplex system can be found from the solution for the frequency components of (B). An analysis shows that the spectrum consists of the applied audio-modulation frequency f_0 and a large number of sidebands $(f_c \pm f_0)$, $(2f_c \pm f_0)(3f_c \pm f_0) \cdots (nf_c \pm f_0)$ in which f_c is the fundamental frequency (line frequency) of the pulse wave.[5] The amplitude of each group obeys the damped sine-wave law $\sin n\pi r/n$ where n is the order of the sideband and r is the ratio of the width of the pulse to the fundamental period. In a television application, r could not exceed about 0.06 and hence the amplitude factor, $\sin n\pi r/n$ changes slowly. When f_0 exceeds $1/2\, f_c$, there is overlapping of the first-order lower sideband and the audio-frequency component, as well as general overlapping of adjacent sidebands of higher order. We do not have knowledge of any detector whereby an undistorted audio signal can be recovered from this multiplicity of overlapping sidebands. However, if the frequency of the audio modulation is restricted by a low-pass filter at the transmitter to less than one half the fundamental pulse frequency, this confusion is avoided. A similar filter must be installed in the receiver for the rejection of frequencies above $f_c/2$. Such a low-pass filter in the receiver functions as a distortionless detector of the audio modulation. Therefore, the theoretical upper limit of the audio bandwidth is equal to one half of line frequency, or 7875 cycles with the present standards.

Vertically scanned pictures would allow a greater maximum audio frequency of $4/3 \times 7875$, or 10,500 cycles.[6] However, it has been observed in laboratory tests that moving subjects scanned vertically with an interlaced pattern do not, in general, reproduce with as much detail as horizontally scanned subjects. This is due, probably, to the predominance of horizontal motion in average subject matter. Hence, it appears that vertical scanning must be rejected as a means of increasing the maximum audio frequency.

The most promising way of increasing the upper audio limit in a monochrome system is an increase in the video bandwidth. The two quantities are related by the formula[7]

$$f_a = K_1\sqrt{f_v} \tag{1}$$

in which

f_a = maximum audio frequency
f_v = video bandwidth
K_1 = a constant.

Fig. 3 shows the correlation between sound band, video band, and number of lines for monochrome and color transmissions. The latter is assumed to be a sequential tricolor system with an interlace ratio of 2:1 and a color-field frequency of 120 cycles.[8] For a given video bandwidth, the quality of duplex sound in terms

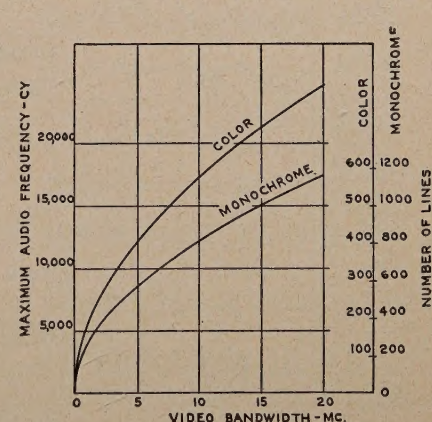


Fig. 3—Maximum audio frequency versus video bandwidth and number of lines.

of audio bandwidth may be made 1.4 times better for color than for monochrome television. Thus, high-fidelity sound (11,000 cycles) may be duplexed along with a color picture of about 360 lines over a video channel of 4 megacycles, while a monochrome picture of 525 lines, which occupies the same video band, accommodates only about 7800 cycles. A maximum audio frequency of 11,000 cycles would require over 700 lines in a monochrome system.

The corresponding radio-frequency channel in all cases is approximately 30 per cent greater than the video bandwidth as a consequence of the additional space required by the vestigial sideband.

A proposal for sound transmission has been disclosed which removes the limitation on the maximum audio frequency of one-half line frequency.[9] In effect, the

[5] Appendix I.
[6] The line frequency in a vertical scanning system is 4/3 the line frequency of the standard horizontal scanning system with an aspect ratio of 4 to 3.
[7] Appendix II.
[8] P. C. Goldmark, J. N. Dyer, E. R. Piore, and J. M. Hollywood, "Color television," Proc. I.R.E., vol. 30, pp. 162–182; April, 1942.
[9] A. V. Bedford, U. S. Patent No. 2,089,639.

system provides for the modulation of a rectangular pulse carrier of two times line frequency and the subsequent delay of alternate pulses to a time position which permits transmission of pairs of pulses during horizontal blanking time. At the receiving point, the previously undelayed pulses are delayed before detection, thus restoring the modulated pulse signal to its original form as a wave of double line frequency. The maximum audio frequency has thereby been increased to line frequency. This system, however, does not appear to be economically feasible from the point of view of receiver design at the present time.

Success or failure of a method of transmission often rests on the degree of immunity to noise. The signal-to-noise ratios appearing in Table I provide a direct comparison of amplitude-modulation pulse transmission and other systems. Comments on the significance of the ratios, and the bearing on modulated pulses as an audio service for television are made later. It is clear that the amplitude-modulation pulses suffer a disadvantage in that superimposed noise cannot be reduced by amplitude limiting to the extent possible in certain other systems.

TABLE I

SIGNAL-TO-NOISE RATIOS (root-mean-square)

Method of Transmission	(1) Signal-to-Noise	(2) Signal-to-Noise (Critical)	(3) Signal-to-Noise $P/N=K/d^2$	(4) Signal-to-Noise $P/N=K/d^2$	(5) Signal-to-Noise (Critical)
Standard Amplitude Modulation	$\frac{2P}{\sqrt{f_a}\,N}$	None	$\frac{2}{\sqrt{f_a}}\frac{K}{d^2}$	$0.023\frac{K}{d^2}$	None
Standard Frequency Modulation	$\frac{\sqrt{3}f_d}{f_a^{3/2}}\cdot\frac{P}{N}$	$\sqrt{b}\left(\frac{f_d}{f_a}\right)^{3/2}$	$\frac{\sqrt{3}f_d}{f_a^{3/2}}\frac{K}{d^2}$	$f_d=150$ kilocycles $0.417\frac{K}{d^2}$	$f_d=150$ kilocycles 217
Amplitude-Modulated Pulses	$\frac{3\sqrt{2}\sqrt{r}}{4\sqrt{f_a}}\frac{P}{N}$	None	$\frac{3\sqrt{2}\sqrt{r}}{4/\sqrt{f_a}}\frac{K}{d^2}$	$0.003\frac{K}{d^2}$	None
Symmetrical Width-Modulated Pulses	$\frac{3w}{2t_s\sqrt{f_v}}\frac{P}{N}$	$\frac{4w}{t_s}$	$\frac{3w}{2t_s\sqrt{f_v}}\frac{K}{d^2}$	$0.012\frac{K}{d^2}$	62
Dissymetrical Width-Modulated Pulses	$\frac{3w}{2t_s\sqrt{f_v}}\frac{P}{N}$	$\frac{4\sqrt{2}w}{t_s}$	$\frac{3w}{\sqrt{2}t_s\sqrt{f_v}}\frac{K}{d^2}$	$0.016\frac{K}{d^2}$	87
Pulses of Frequency Modulation	$\frac{3\sqrt{6}\sqrt{r}f_d}{8f_a^{3/2}}\frac{P}{N}$	$\sqrt{6}\sqrt{r}\left(\frac{f_d}{f_a}\right)^{3/2}$	$\frac{3\sqrt{6}\sqrt{r}f_d}{8f_a^{3/2}}\frac{K}{d^2}$	$f_d=150$ kilocycles $0.052\frac{K}{d^2}$	$f_d=150$ kilocycles 107
Pulses of Frequency Modulation	$\frac{3\sqrt{6}\sqrt{r}f_d}{8f_a^{3/2}}\frac{P}{N}$	$\sqrt{6}\sqrt{r}\left(\frac{f_d}{f_a}\right)^{3/2}$	$\frac{3\sqrt{6}\sqrt{r}f_d}{8f_a^{3/2}}\frac{K}{d^2}$	$f_d=1200$ kilocycles $0.417\frac{K}{d^2}$	$f_d=1200$ kilocycles 2420

The signal-to-noise formulas in columns (1), (3), and (4) of the table for standard frequency modulation, width-modulated pulses, and pulses of frequency modulation during postblanking, are valid only for ratios higher than the critical ratio since the formulas are derived with the assumption that noise limiting is effective. Thus it may appear, with only a casual reading of the table, that the ratio is always the same for standard frequency modulation and pulses of frequency modulation $f_d=1200$ kilocycles. The fact is that the two types of transmission yield equal signal-to-noise ratios only when the critical ratio for pulses of frequency modulation is exceeded. At greater distances from the transmitter standard frequency modulation is superior.

Values of constants: $f_a=7500$ cycles per second; $r=0.06$; $w/t_s=15.4$; $f_v=4\times10^6$ cycles per second.

2. *Width-Modulated Rectangular Pulses*

A more promising form of pulse modulation, from the standpoint of signal-to-noise ratio, is a constant-amplitude pulse system wherein the width of a pulse is proportional to the amplitude of the audio signal. Two examples of width-modulated signals are illustrated in Fig. 4(B) and (C). In type (1), a dissymmetrical modulation, the leading edges of the pulses occur periodically, but the widths are proportional to the instantaneous

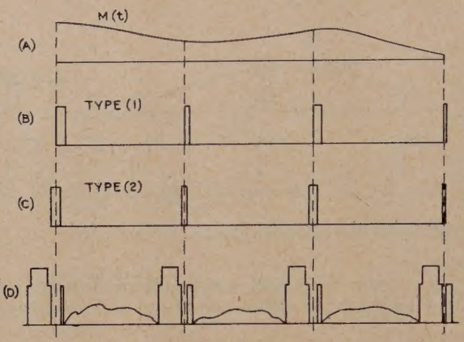


Fig. 4—Width-modulated duplex wave forms.

(A) audio modulation $M(t)$
(B) pulse carrier dissymmetrically width-modulated by $M(t)$—type (1)
(C) pulse carrier symmetrically width-modulated by $M(t)$—type (2)
(D) combination of type (2) and television wave form

amplitude of the audio signal $M(t)$ at the instant of the leading edge· that is, only the trailing edge is "modulated." Type (2) is a symmetrical modulation, the width of a pulse being proportional to the instantaneous amplitude at the instant corresponding to the center line of the unmodulated pulse. The center lines are periodically spaced; thus, both leading and trailing edges of type (2) are modulated. Such pulse waves may be inverted in polarity and combined with the standard television wave form as shown in Fig. 4(D) for type (2) modulation. Any amplitude variation of the pulse due to noise may be removed by limiting in the receiver following separation of the pulse from the video signal if the peak noise does not exceed one half of the pulse amplitude.

Equation (28) in Appendix III is the expression for a pulse wave width-modulated in the symmetrical manner (type 2) by a sine wave. The similarity to standard frequency modulation is noticeable in the sequence of sidebands which are generated. Thus when a pulse carrier of fundamental frequency f_c is width-modulated at a rate of f_0 cycles per second, the resultant wave contains component frequencies f_c, (f_c+f_0), (f_c-f_0), (f_c+2f_0), (f_c-2f_0), etc., as well as corresponding sidebands for each harmonic of the fundamental f_c; namely $2f_c$, $(2f_c+f_0)$, $(2f_c-f_0)$, $(2f_c+2f_0)$, $(2f_c-2f_0)$, etc. In addition, the frequency terms containing only the modulating frequency f_0 and its harmonics $2f_0$, $3f_0$, etc. appear. The general term is $(Mf_c \pm Nf_0)$ where M and N are positive integers or zero.

Since overlapping of f_0 and the sideband (f_c-f_0) must be prevented, the modulating frequency should not exceed $f_c/2$. At the receiving end, the modulated pulse-signal may be applied to a low-pass filter which rejects all sidebands and harmonics exceeding one half the

frequency of the fundamental f_c. However, all distortion terms are not thereby excluded; harmonics of f_0 and the sidebands of higher order (f_c-2f_0), (f_c-3f_0), may fall within the pass band. The magnitudes of the most important distortion terms in (28) have been plotted in Fig. 5. The modulation constant α was taken equal to 1, the value corresponding to maximum modulation. A value of 3 per cent was assigned to w, the unmodulated pulse width. Therefore, the widths of the pulses in the modulated wave vary from 0 to 6 per cent of the period of the carrier. This is substantially the maximum variation under the specifications given for the television synchronizing wave form (Fig. 1). The broken-line portion of a curve indicates the range of the component which is suppressed by the low-pass filter in the receiver. Thus the component (f_c-2f_0) is suppressed for the range $f_0<f_c/4$ but is transmitted when $f_0>f_c/4$. In a reverse manner the component $2f_0$ is transmitted when $f_0<f_c/4$ and suppressed when $f_0>f_c/4$. The maximum value attained by either component is approximately 0.05 per cent of the amplitude of the audio component f_0.

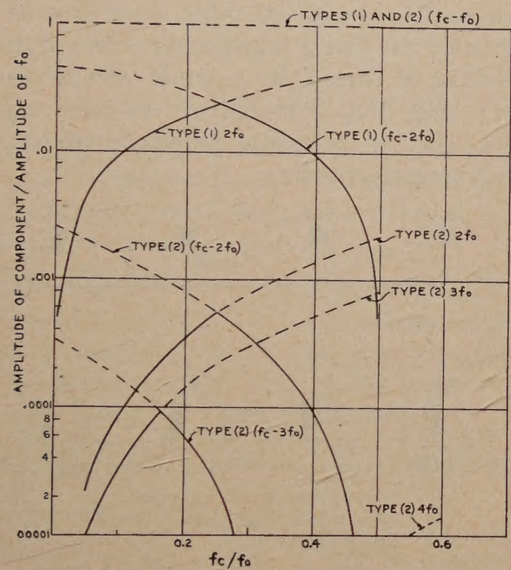


Fig 5—Frequency components resulting from width modulation of a pulse carrier wave by a sine wave.
Type (1) = dissymmetrical modulation
Type (2) = symmetrical modulation
f_c = pulse frequency
f_0 = modulating frequency

An analysis of a dissymmetrical width-modulated pulse wave (Appendix III, equation (31)) displays the same general characteristics as the symmetrical modulation. Harmonics of the audio frequency, as well as numerous sidebands, are present, but the magnitudes shown in Fig. 5 are greater than in type (2) modulation. The largest contribution of any distortion term is 2.5 per cent.

Signal-to-noise ratios calculated according to the derivations in Appendix IV appear in Table I and Fig. 15.

Fig. 6(A) illustrates one method for the production of dissymmetrical width-modulated pulses. The starting point is a wave of narrow triangular pulses (Fig. 6(B)) which is derived from driving pulses at line frequency normally generated by the synchronizing generator. To the triangular pulses is added the audio signal from which the frequency components higher than one half the line frequency have been removed by a low-pass filter. Limiter No. 1 removes the audio wave below the base line as shown in Fig. 6(B). The residue is greatly amplified and then acted upon by limiter No. 2, with the result that width-modulated pulses of substantially rectangular shape are produced. These are inserted in the line-blanking interval following the synchronizing pulse (postblanking). The standard field synchronizing pulse must be slotted down to black level during the

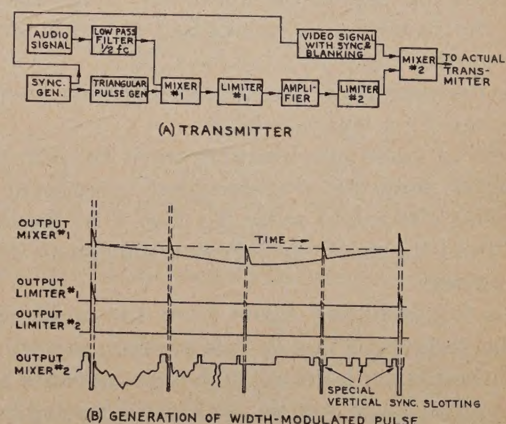


Fig 6—Width-modulated duplex-system transmitter.

line pulses, as shown in Fig. 6(B), in order that the width-modulated pulses when applied may extend to white level. The combined video signal is applied to the picture transmitter in the customary manner.

Fig. 7(A) illustrates the functional arrangement of the receiver. The selection of the width-modulated pulses

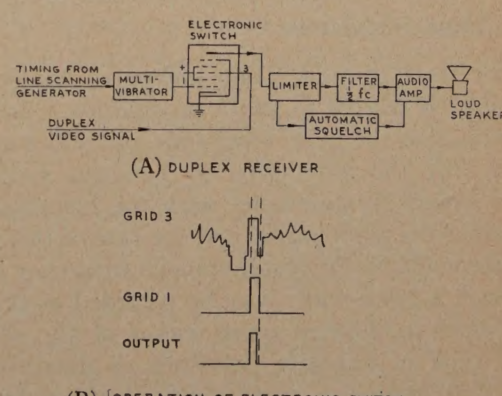


Fig 7—Width-modulated duplex system. Type (1) receiver.

from the video signal and rejection of picture components is performed by an electronic switch, usually a vacuum tube having two control grids. A keying pulse originating in a multivibrator which is synchronized by

the line-scanning circuit is impressed on grid 1 of the switch. The width and timing of the pulse is critical for the most favorable signal-to-noise ratio. The duplex video applied to grid 3 causes plate current to flow only when the tube is keyed on. In this way the width-modulated pulses are isolated.

Amplitude noise is removed by limiting if the peak noise does not exceed one half of the pulse amplitude but the variation of the pulse width due to noise is not removable. Such variation constitutes a width modulation and is reproduced as audible noise. When the synchronization of the receiver is impaired by noise, the timing of the switch is likewise affected, and parts of blanking and picture may be admitted into the audio amplifier and appear as noise in the loud speaker. There is a marked increase in the immunity of the system to noise if the receiver is synchronized by automatic frequency control.[10]

Audio components in excess of one-half line frequency are removed by a low-pass filter.

Additional kinescope blanking must be provided in the receiver since the duplex signal extends to white level during the sound pulse. In Fig. 7(A) blanking is derived from the multivibrator simultaneously with the keying pulses.

Means for excluding signal from the audio circuits when the receiver is not in synchronism is very desirable. Without such a device, video components are admitted to the audio system with an annoying audible result. A circuit may be devised which is sensitive to the changed character of the signal passed by the electronic switch during intervals of missynchronization and applies a bias beyond cutoff to the audio amplifier.

In September, 1943, television signals containing width-modulated pulses of the dissymmetrical type were transmitted by television station WNBT, and successfully received in Princeton, New Jersey, using the system outlined above.

3. *Pulse Time Modulation*

Another form of modulation known as "pulse time modulation" is related to width modulation.[11] In pulse time modulation the pulse amplitude and width remain constant, but the time interval between successive pulses is varied in accordance with the instantaneous amplitude of the audio signal and the rate of this variation corresponds to the instantaneous frequency of the signal. Such a pulse wave may be regarded as the sum of two width-modulated pulse waves of the dissymmetrical type of opposite polarities as illustrated in Fig. 17. The frequency components of the pulse time wave are therefore solvable from (31).

[10] K. R. Wendt and G. L. Fredendall, "Automatic frequency and phase control of synchronization in television receivers," Proc. I.R.E., vol. 31, pp. 7–15; January, 1943.

[11] E. M. Deloraine and Emil Labin, "Pulse time modulation," *Elec. Commun.*, vol. 22, pp. 91–98; 1944.

4. *Pulsed Frequency Modulation*

In contrast with the foregoing duplex methods involving rectangular pulses for the transmission of sound during the line-blanking interval, there is a method which may be called "pulsed frequency modulation," that employs wave bursts of a frequency-modulated subcarrier for the same purpose. The bursts are generated at the transmitter by a sine-wave oscillator which is operative only during line blanking and is frequency modulated by the audio signal during this interval. The center frequency and deviation are chosen so that the essential sidebands lie within the video band. These subcarrier bursts are combined with the video wave form as modulation of either the line synchronizing pulses or the post-blanking (Fig. 8). From the point of view of signal-to-noise ratio, Fig. 8(B) is preferable. In either case, the composite signal is applied as amplitude modulation of the radio-frequency carrier.

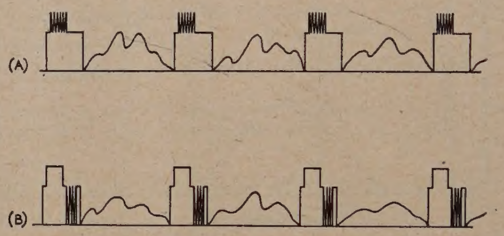


Fig. 8—Pulsed frequency-modulation duplex wave forms.
(A) in synchronizing pulse
(B) in post blanking

In the receiver, the bursts are first isolated at the video level from the picture part of the composite wave, then amplitude limited for removal of noise, and finally applied to a conventional balanced frequency-modulation discriminator centered at the frequency of the subcarrier. The output of the discriminator is an amplitude-modulated pulse wave. The audio signal is derived from the pulse output of the discriminator by removing all components in excess of one-half line frequency with a low-pass filter.

As the result of experimental and theoretical work with pulsed frequency modulation, certain features of the technique were discovered which could escape a casual study. Such matters are treated in the following discussion.

A. Transient response of pulsed frequency-modulated circuits: If a pulsed frequency-modulated system is to function properly, the peak value of the detected pulses should depend solely on the instantaneous frequency of the subcarrier. This means that the various tuned circuits involved should complete their transients in a time which is short compared with the total duration of the wave burst. Fig. 9 shows the response of a simple tuned circuit to a wave burst of constant amplitude that starts and stops with zero phase and lasts T_p seconds. The circuit has a build-up time τ, which may be adjusted by the damping resistor R and is correlated with the

bandwidth b in the form

$$\tau = 2RC = 1/\pi b. \tag{2}$$

The minimum bandwidth is determined by the time allowed for the transients. If these are to be complete within p percent of the pulse time,

$$b \geqq \frac{100}{\pi p T_p}. \tag{3}$$

The total number n of cycles per pulse, as well as the number Δn occurring before the steady state is attained, are

$$n = T_p f_s \tag{4}$$

$$\Delta n = \tau f_s. \tag{5}$$

Equation (5) holds regardless of the subcarrier frequency f_s. The following set of constants is representative of a typical circuit designed for pulsed frequency-modulation operation:

Pulse time T	5 microseconds
Time of build-up τ	0.5 microsecond
Circuit capacitance C	25 micromicrofarads
Circuit resistance R	10,000 ohms
Bandwidth b	400 kilocycles
Subcarrier frequency f_s	4 megacycles
Q factor	10
Cycles per pulse n	20
Δn	2 cycles

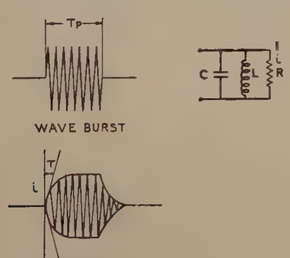


Fig. 9—Response of a tuned circuit to a pulse of frequency modulation.

B. Generation of phasing of the subcarrier at the transmitter: According to the calculation above, a total variation of no more than 2 subcarrier cycles is sufficient to produce peak modulation in the receiver. Hence it follows that the instantaneous wave forms of the subcarrier bursts must be closely similar at the beginning. Unless the initial phase of each burst is repeated with extreme accuracy, the otherwise random initial phases may introduce audible beat notes and noise in the detected signal.

In this connection, the keying of the subcarrier for part-time modulation presents a major problem. If an independent subcarrier oscillator supplying a continuous frequency-modulated wave is used, an electronic on-off switch controlled by the main synchronizing generator must be provided. This switch cuts into the subcarrier and admits sections of its wave train for modulation of the synchronizing pulses (or blanking). It is obvious that the timing of this switch would have to be accurate within small fractions of one subcarrier cycle, or about 0.1 microsecond, in order that the initial phases of all pulses be substantially equal. The noise susceptibility of this method is high, because the subcarrier modulation is keyed on and off at full amplitude.

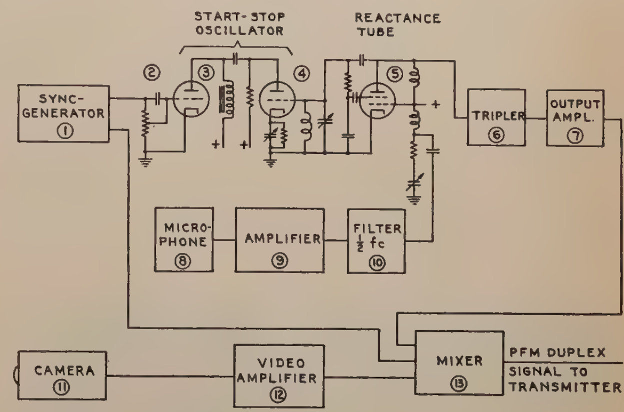


Fig. 10—Transmitter for pulsed frequency modulation.

The problem of precise keying is further aggravated by the fact that the repetition frequency of television pulses is not constant. In all practical television synchronizing generators, the line frequency is subjected to continuous frequency control so that it constitutes, at any instant, a definite multiple of the field frequency. The field frequency is synchronized with a 60-cycle power supply which is inherently variable around a well-defined average. As a result, beat notes of variable pitch are bound to occur if a subcarrier source with continuous frequency modulation and constant center frequency is subjected to keying from a synchronizing generator unless special precautions are taken.

In the system described below, such spurious signals have been effectively eliminated. A continuous subcarrier generator is not used; instead, the bursts are supplied from a start-stop oscillator which is switched on and off by the line blanking pulses. The start-stop subcarrier oscillator shown at (4) in Fig. 10 is active only when plate voltage is applied in the form of a pulse from the control tube (3). Pulses of appropriate wave shape at line frequency are derived directly from the synchronizing generator and impressed on the grid of the control tube. Hence throughout the line-scanning interval the subcarrier oscillator is inoperative, but at the end of each line it receives a plate-voltage pulse. As a result, subcarrier oscillations are built up with exactly the same initial phase conditions each time. Since the plate-power pulse is derived from the line-blanking pulse, it participates automatically in any variations of the line frequency. The power pulses may also be preshaped in such a manner that the plate supply ceases in time to allow

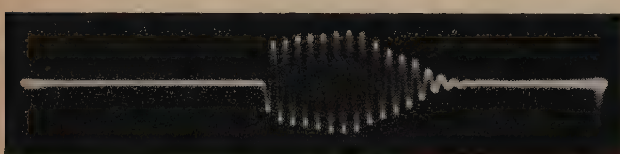
the subcarrier oscillations to decay within the allotted duration of sound transmission.

Fig. 11 illustrates the subcarrier burst without and with frequency modulation.

In Fig. 11(B), which shows a large number of frequency-modulated pulses in superposition, the first half of the wave burst is sharp while the wave trace appears increasingly blurred toward the end. This verifies the fact that the initial phase is substantially identical for all bursts regardless of the frequency modulation: that is, the pulse fronts are "coherent."

C. Pulsed frequency-modulation transmitter: Fig. 10 shows a possible arrangement of components in a pulsed frequency-modulation transmitter. The start-stop oscillator is coupled to a reactance tube (5) which is controlled continuously by the audio signal. A low-pass filter (10) with a cutoff at one half of the line frequency prevents the generation of overlapping sidebands of the

(A)



(B)

Fig. 11—(A) subcarrier burst
(B) subcarrier burst, frequency-modulated

subcarrier that would interfere with audio fidelity. In an experimental transmitter, the master oscillator generated about 10 cycles at a frequency of 2 megacycles during each burst with a deviation of ± 100 kilocycles. At the output of the doubler stage (6) the center frequency becomes 4 megacycles and the deviation ± 200 kilocycles. The subcarrier burst is amplified and combined with the video signal at (13).

From the point of view of pulsed frequency modulation the field synchronizing pulse and the equalizing pulses interfere with the regular sequence of horizontal synchronizing pulses. Some modification of the standard television wave form (Fig. 1) is necessary to allow the transmission of wave bursts of constant duration. Interruptions in the sequence result in the generation of a narrow 60-cycle pulse that contains harmonics of 60 cycles extending throughout the audible spectrum.

If the bursts occur during postblanking (Fig. 8(B)) the field-synchronizing pulse may be slotted as in Fig. 6(B), but if the line-synchronizing pulses are modulated by the bursts, the modification shown in Fig. 12 is desirable. Here the slots S_1 isolate the subcarrier bursts from the field signal so that separation of the sound may take place in the receiver. The slots S_2 act as equalizers for maintenance of interlacing. Fig. 13 shows the modified television wave form carrying pulsed frequency-modulation duplex on the line synchronizing.

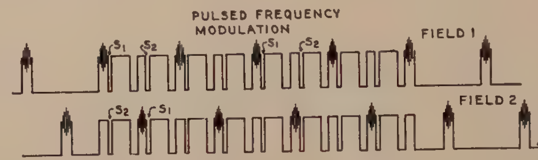


Fig. 12—Modification of television wave form for pulsed frequency modulation on line-synchronizing pulses.

D. Pulsed frequency-modulation receiver: A complete pulsed frequency-modulation receiver is shown in Fig. 14. Isolation of the frequency-modulation bursts (whether

Fig. 13—Combined video signal and pulsed frequency modulation of line-synchronizing pulses.

in line-synchronizing pulses as in Fig. 8(A) or in postblanking, as in Fig. 8(B)) is performed by a selector such as a tube with two control grids. The selector is biased off by a suitable pulse signal generated by a multivibrator which is synchronized from the line-

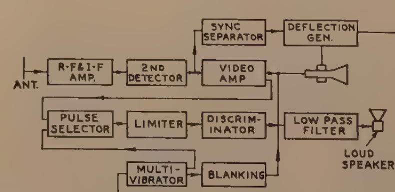


Fig. 14—Pulsed frequency-modulation receiver.

deflection generator or the line-synchronizing circuits of the video receiver. A limiter removes the amplitude noise to substantially the same extent as in conventional frequency-modulation systems. Demodulation of the bursts is accomplished in a conventional discriminator circuit centered at the subcarrier frequency. All audio components above a frequency of one-half line frequency are removed by a low-pass filter as in the other duplex

systems mentioned above. A locally generated blanking signal is required for biasing off the kinescope when the wave form of Fig. 8(B) is used.

Signal-to-Noise Ratios of Duplex and Standard Systems

Formulas for the signal-to-noise ratios of duplex and standard systems are derived in Appendix IV. A comparison of the various ratios requires the assumption of a numerical relationship between the amplitudes of the respective carriers.

The usual practice in television installations is to establish the amplitude S of the standard frequency-modulated sound carrier at about 0.7 of the peak amplitude P of the picture carrier. For convenience the ratio S/P will be taken as $1/\sqrt{2}$. When amplitude modulation was standard for sound transmission prior to the adoption of frequency modulation, the same ratio $1/\sqrt{2}$ was customary.

The amplitudes in duplex transmission are fixed by the amplitude of the picture carrier.

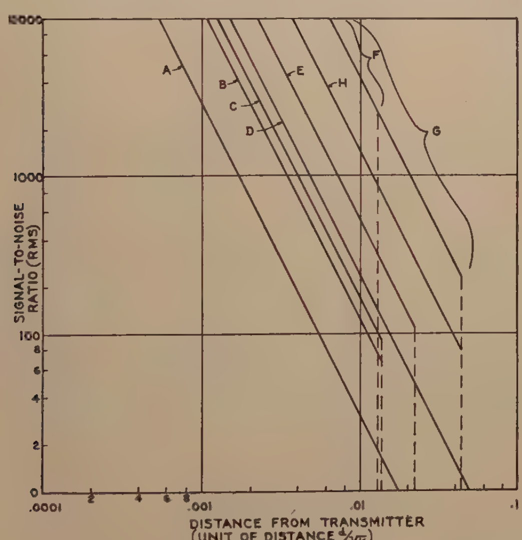


Fig. 15—Signal-to-noise ratios for sound transmission.
A = amplitude-modulated pulses
B = width-modulated pulses (symmetrical)
C = width-modulated pulses (dissymmetrical)
D = standard amplitude modulation
E = pulsed frequency modulation (f_d = 150 kilocycles)
F = pulsed frequency modulation (f_d = 1200 kilocycles)
G = standard frequency modulation (f_d = 150 kilocycles)
H = standard frequency modulation (f_d = 50 kilocycles; bandwidth = 150 kilocycles).

The unmodulated amplitude h of the amplitude-modulated pulse carrier is one half the amplitude of blanking. In the standard wave form (Fig. 1) blanking is three fourths of the peak amplitude of the picture carrier. Hence h may be taken as $3P/8$. The amplitude of the pulsed frequency-modulation signal during postblanking is also $3P/8$. The amplitude H of width-modulated pulses is equal to the full amplitude of blanking or $3P/4$.

Column (1) of Table I lists the audio signal-to-noise ratios for the various methods of transmission of sound in terms of P/N and other dimensions which are associated with a particular method. P is the amplitude of the picture carrier and N is the noise factor.

Column (2) lists the critical signal-to-noise ratios below which the formulas are no longer valid. This limit exists in the case of width-modulated pulses when the peak noise is higher than one half the pulse amplitude. There is no limit in standard amplitude modulation and amplitude-modulated pulses since limiting is not applied. The limit occurs in standard frequency modulation and pulsed frequency modulation when the peak amplitudes of noise and signal are equal.

If the noise is assumed to remain constant, the signal, and therefore the signal-to-noise ratio, varies with distance from the transmitter according to the law for the propagation of television signals. Hence P/N may be replaced by K/d^2 as shown in column (3) where d is the distance from the transmitter and K is a proportionality constant. If the distance d exceeds the line-of-sight distance, a somewhat higher power of d would be appropriate.

Columns (4) and (5) show the forms taken by (2) and (3) when values are substituted.

Fig. 15 illustrates the variation of the signal-to-noise ratios with distance from the transmitter. Comparisons made of the various methods of sound transmission from Fig. 15 are necessarily on a relative basis since the unit of distance is $d/\sqrt{K}$.

Standard frequency modulation with a deviation of 150 kilocycles yields the most favorable signal-to-noise ratio within 0.044 units of distance. An equal ratio may be obtained over a more limited distance of 0.013 units with pulsed frequency modulation during postblanking if the maximum deviation is of the order of 1.2 megacycles. A greater deviation is required in pulsed frequency modulation for equality, because the audio signal which may be recovered from a pulsed-frequency-modulation wave is proportional to the pulse width, whereas the audio noise is proportional to the square root of the width.[12] The maximum distance from the transmitter at which limiting of a pulsed frequency-modulation signal is effective in removing noise (that is, the critical distance) is necessarily less because the noise voltage admitted to the receiver is greater as a consequence of the greater deviation.

Pulsed frequency modulation during postblanking with the customary deviation of 150 kilocycles is intermediate between standard amplitude modulation and standard frequency modulation.

Width-modulated pulses are intermediate between amplitude-modulation pulses and standard amplitude modulation.

Amplitude-modulated pulses rank lowest, largely as a consequence of not being susceptible to limiting.

[12] Appendix IV, equation (43).

Other Receiver Considerations

A duplex receiver is "no better" than its sound pulse selector. Audible noise can be introduced into the audio system of a duplex receiver when portions of the video signal, representing picture, are selected along with the desired sound signal. This occurs when the accuracy of synchronization of the selector is reduced sufficiently by noise and interference. In this respect, the automatic frequency control of synchronization was found to be definitely superior to conventional triggered synchronization.[10] The flywheel effect of the automatic-frequency-control circuit tends to minimize the disturbing effect of noise on synchronization.

It appears that with automatic-frequency-control synchronization the major part of the total audible noise in a well-designed duplex system may be attributed to the inherent noise characteristics discussed in Appendix IV rather than to inaccurate selection of the sound signal.

The stability of duplex circuits was not studied, but it is clear that drifts in the values of circuit elements that affect the accuracy of sound selection would be detrimental.

An exhaustive study of the relative costs of a television receiver designed for duplex sound on a conventional receiver intended for reception of standard frequency modulation was not included in the scope of this project. However, an analysis of two experimental receivers constructed according to the arrangements in Figs. 7(A) and 14 indicates that the cost of a commercial duplex receiver is not likely to exceed that of a standard receiver.

Appendix I

A rectangular-pulse wave of unit amplitude may be expressed as a cosine series

$$e(t) = r + \frac{2}{\pi}\left\{\sin \pi r \cos \omega_c t + \frac{\sin 2\pi r}{2} \cos 2\omega_c t + \cdots + \frac{\sin n\pi r}{n} \cos n\omega_c t + \cdots\right\}$$

$$\left\{\begin{array}{l}\omega_c = 2\pi f_c \\ r = \text{pulse width per pulse period.}\end{array}\right. \tag{6}$$

Modulation of $e(t)$ by an audio signal $M(t)$ has the result

$$f(t) = [1 + M(t)]e(t)$$

$$= \underbrace{e(t)}_{\text{original unmodulated pulse wave}} + \underbrace{rM(t)}_{\text{audio signal diminished by } r} + \underbrace{\sum_{n=1}^{\infty} \frac{2}{\pi} M(t) \frac{\sin n\pi r}{n} \cos n\omega_c t}_{\text{sidebands of carrier and its harmonics}}. \tag{7}$$

Appendix II

A well-known formula[13] expressing the video bandwidth required for equal horizontal and vertical resolution is

$$f_v = \tfrac{1}{2}KL^2Na \tag{8}$$

in which

f_v = video bandwidth
L = number of scanning lines
N = frame repetition rate
a = aspect ratio
K = experimental factor often taken equal to 0.6.

Since the maximum audio frequency f_a which may be transmitted by a pulse carrier is $LN/2$, the combination of this formula and (8) gives

$$f_a = \sqrt{\frac{f_v N}{2K_a}} = K_1\sqrt{f_v}. \tag{9}$$

Equation (9) should be regarded chiefly as an expression of proportionality between the quantities because the value of K depends upon the criterion for equal resolutions, which is not a precise concept.

Appendix III

Frequency Components Resulting from Symmetrical Width Modulation of a Rectangular-Pulse Carrier by a Sine Wave

The problem is the calculation of the amplitude and frequency of each component in a rectangular-pulse carrier which is width-modulated in a symmetrical manner by a sine wave. The width of a pulse is proportional to the amplitude of the modulating wave at the instant corresponding to the center line of the pulse. Hence, the width a_p of the pth pulse is

$$a_p = w\left(1 - \propto \cos 2\pi p \frac{T_c}{T_o}\right) \tag{10}$$

in which

w = width of unmodulated pulse
$T_c = 1/f_c$ = period of pulse wave
$T_o = 1/f_0$ = period of modulating wave
$\propto$ = modulation factor.

In the general case, the modulated carrier wave will not repeat precisely at the end of each audio cycle, but after some time greater than the period T_0 there will be repetition. Let this time be called T and the corresponding frequency, f. The equation of the modulated pulse wave may be deduced by regarding the wave as the summation of a large number of pulse waves of equal period T. Each component wave will be characterized by a certain pulse width which is constant for the particular component. Thus there is a wave starting at the origin and characterized by a pulse width a_0, a wave of

[13] R. D. Kell, A. V. Bedford, and M. A. Trainer, "An experimental television system, Part II. The transmitter," Proc. I.R.E., vol. 22, pp. 1246–1266; November, 1934.

width a_1 and phase T_c, a wave of width a_2 and phase $2T_c$, etc.

The pth wave has a width a_p and phase pT_c. There are (f_cT-1) waves to sum. The equation of the pth wave is

$$e_p = \left[\frac{a_p}{T} + \frac{2}{\pi}\sum_{n=1}^{\infty}\frac{\sin n\pi a_p f}{n}\cos 2\pi n f(t - pT_c)\right]. \quad (11)$$

A summation over p yields the equation of the modulated pulse wave.

$$e' = \sum_{p=0}^{f_cT-1}\frac{a_p}{T} + \frac{2}{\pi}\sum_{p=0}^{f_cT-1}\sum_{n=1}^{\infty}\frac{\sin n\pi a_p f}{n}\cos 2\pi n f(t-pT_c)$$
$$= e_{\text{d.c.}} + e. \quad (12)$$

Since the direct-current component is not of interest, it need not be considered further. If a certain frequency component of e is sought, the contributions of each of the p waves must be summed in the form

$$\frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{\sin n\pi a_p f}{n}\cos 2\pi f n(t - pT_c). \quad (13)$$

The amplitude of the component (fn) is $\sqrt{A_1^2+B_1^2}$ in which

$$A_1 = \frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{\sin (n\pi a_p f)}{n}\cos 2\pi f n p T_c \quad (14)$$

$$B_1 = \frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{\sin (n\pi a_p f)}{n}\sin 2\pi f n p T_c. \quad (15)$$

The remainder of the derivation is devoted to an examination of A_1 and B_1. It is expected that only certain values of n will lead to nonzero values for A_1 and B_1. Before summation, the expression for a_p is inserted in A_1 and B_1. There results

$$A_1 = \frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{1}{n}\sin\left[2n\pi f\epsilon(1 - \propto \cos 2\pi T_c f_0 p)\right] \cos 2\pi f n p T_c \quad (16)$$

$$B_1 = \frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{1}{n}\sin\left[2n\pi f\epsilon(1 - \propto \cos 2\pi T_c f_0 p)\right] \sin 2\pi f n p T_c. \quad (17)$$

Expansion of A_1, in terms of Bessel functions, yields

$$A_1 = \frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{1}{n}\left[\cos Rp \sin \sum_{s\,(\text{even})=2}^{\infty} 2(-1)^{s/2}J_s(B)\cos sC_p + J_0(B) + \cos Rp \cos\cdot A \sum_{v\,(\text{odd})=1}^{\infty} 2(-1)^{(v+1)/2}J_v(B)\cos vC_p\right] \quad (18)$$

and

$$B_1 = \frac{2}{\pi}\sum_{p=0}^{f_cT-1}\frac{1}{n}\left[\sin Rp \sin A \sum_{s\,(\text{even})=2}^{\infty} 2(-1)^{s/2}J_s(B)\cos sC_p + \sin Rp\cos A \sum_{v\,(\text{odd})=1}^{\infty} 2(-1)^{(v+1)/2}J_v \cos vC_p\right] \quad (19)$$

in which

$$\pi n f w = A \qquad 2\pi T_c f_c = C$$
$$n\pi w \propto f = B \qquad 2\pi f n T_c = R. \quad (20)$$

A typical term in A_1 involving the summation over s is

$$\frac{4(-1)^{s/2}}{\pi n}J_s \sin A \sum_{p=0}^{f_cT-1}\cos Rp \cos sCp. \quad (21)$$

The expression

$$\sum_{p=0}^{f_cT-1}\cos Rp \cos sCp \quad (22)$$

is a finite trigonometric sum which is known to have the value

$$\frac{1}{2}\,\frac{\cos[(f_cT-1)(R-sC)/2]\sin[f_cT(R-sC)/2]}{\sin[(R-sC)/2]} + \frac{1}{2}\,\frac{\cos[(f_cT-1)(R+sC)/2]\sin[f_cT(R+sC)/2]}{\sin[(R+sC)/2]}. \quad (23)$$

The above sum may be abbreviated

$$S = \frac{S_1}{2} + \frac{S_2}{2}. \quad (24)$$

If the expressions for R and C in (20) are introduced, S_1 in (24) becomes

$$S_1 = \frac{\cos\pi[sT_cf_0 - nT_cf + n - sTf_0]\sin\pi[n - sTf_0]}{n\sin\pi(nT_cf - sT_cf_0)}. \quad (25)$$

If S_1 is to have a nonzero solution, the denominator must be zero at least for some values of n. This follows from the observation that $\sin\pi(n - sTf_0)$ is always zero. From inspection, it is seen that the denominator of (25) is zero when

$$n = \pm MTf_c + sTf_0 \quad (26)$$

in which M is the positive integer. When (26) is inserted in (25), the indeterminancy may be reduced to

$$S_1 = \frac{1}{sT_cf_0 \pm M}. \quad (27)$$

Similar reasoning leads to explicit forms for A_1 and B_1 in

(14) and (15). Finally, (12), for the modulated wave, may be written in the form

$$e = \frac{2}{\pi} \sum_{M=0}^{\infty} \left[\sum_{\nu=-\infty}^{\infty} \{J_{|\nu|}[\pi w \propto (Mf_c + \nu f_0)]\} \frac{\sin [\pi w(Mf_c + \nu f_0) - |\nu| \pi/2)}{M + \nu f_0/f_c} \cos 2\pi(Mf_c + \nu f_0)t \right]. \quad (28)$$

Frequency Components Resulting from Dissymmetrical Width Modulation of a Rectangular-Pulse Carrier by a Sine Wave

When each pulse of a symmetrically-modulated pulse carrier is translated to the right (or left) on the time axis by an amount equal to one half the width of the modulated pulse, the carrier becomes unsymmetrically modulated. Therefore (12) may be modified to read

$$e' = \sum_{p=0}^{f_cT-1} \frac{a_p}{T} + \frac{2}{\pi} \sum_{p=0}^{f_cT-1} \sum_{n=1}^{\infty} \frac{\sin n\pi a_p f}{n} \cos 2\pi nf \left[t - pT_c - \frac{a_p}{2} \right] \quad (29)$$

in which

$$a_p = w(1 - \propto \cos 2\pi pT_c f_0). \quad (30)$$

A mathematical process similar to that outlined in Appendix II yields the result

$$\begin{aligned} e = &\sum_{v(\text{odd})=1}^{\infty} \frac{1}{\pi} (-1)^{(v+1)/2} \frac{J_v[2\pi w \propto (Mf_c + vf_0)]}{M + vf_0/f_c} \cos 2\pi[(Mf_c + vf_0)(t - w)] \\ &+ \sum_{v(\text{odd})=1}^{\infty} \frac{1}{\pi} (-1)^{(v+1)/2} \frac{J_v[2\pi w \propto (Mf_c - vf_0)]}{M - vf_0/f_c} \cos 2\pi[(Mf_c - vf_0)(t - w)] \\ &+ \frac{2}{\pi} \sum_{\xi}^{\infty} (-1)^{(p+\beta-2)/2} \frac{J_p[\pi w \propto (Mf_c + \xi f_0)] J_\beta[2\pi\epsilon \propto (M + \xi f_0)]}{M + \xi f_0/f_c} \sin 2\pi[(Mf_c + \xi f_0)(t - w) \\ &+ \frac{2}{M\pi} \sin(\pi Mf_c w) \cos 2\pi f_c M(t - w/2) \end{aligned} \quad (31)$$

$$\xi = (\rho + \beta), (\rho - \beta), (-\rho + \beta), (-\rho - \beta)$$

ρ and β are odd positive integers.

Other symbols have the same significance previously assigned.

Appendix IV

Signal-to-Noise Ratios

1. Standard Amplitude Modulation and Standard Frequency Modulation

The root-mean-square signal-to-noise ratio for 100 per cent amplitude modulation is

$$\frac{S}{\sqrt{2}} \Big/ \frac{N\sqrt{f_a}}{4} = 2\sqrt{2} \frac{S}{N\sqrt{f_a}} \quad (32)$$

in which

S = unmodulated amplitude of carrier

f_a = highest audio frequency

$N\sqrt{f_a}$ = peak amplitude of noise (= 4 × root-mean-square noise)[14]

Crosby[15] and others have shown that

$$\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{FM}} = \frac{\sqrt{3} f_d}{2f_a} \left(\frac{\text{signal}}{\text{noise}}\right)_{\text{AM}} \quad (33)$$

in which f_d is 2 times frequency deviation. The amplitude of noise is assumed to be below the threshold value.

From (32) and (33),

$$\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{FM}} = \frac{\sqrt{6} S \sqrt{f_d}}{N f_a} \sqrt{\frac{f_d}{f_a}}. \quad (34)$$

2. Amplitude-Modulated Pulses

In a 100 per cent amplitude-modulated pulse system the root-mean-square value of the audio signal detected by means of a low-pass filter in the receiver (see Appendix I) is

$$\frac{rh}{\sqrt{2}} \quad (35)$$

in which

h = unmodulated height of the pulse

r = ratio of pulse width to period of pulse carrier.

If the assumption is made that the pulse wave is applied to the detector (low-pass filter) only during the time of

[14] Vernon D. Landon, "The distribution of amplitude with time in fluctuation noise," Proc. I.R.E., vol. 29, pp. 50–55; February, 1941.

[15] Murray G. Crosby, "Frequency-modulation noise characteristics," Proc. I.R.E., vol. 25, pp. 472–517; April, 1937.

the pulses, the root-mean-square noise is

$$\sqrt{r}\,\frac{N\sqrt{f_a}}{4}. \qquad (36)$$

Hence

$$\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{AM pulses}} = 2\sqrt{2}\,\frac{h\sqrt{r}}{N\sqrt{f_a}}.$$

3. *Width-Modulated Pulses*

Before detection, the noisy signal is clipped, or limited, top and bottom so that only a comparatively narrow section near the center of each pulse is selected. Hence, it is assumed that noise effects are introduced into width-modulated pulses chiefly by the random displacement of the sides of the pulses. The following derivation applies when the peak noise is less than one half of the amplitude of the pulses. In Fig. 16, let

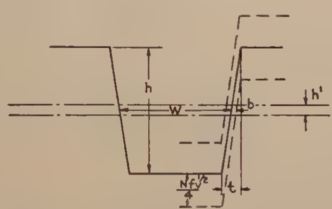


Fig. 16—Width-modulated pulses.

w = unmodulated width of pulse
H = height of pulse
t_s = time of rise of pulse side
b = displacement of side in seconds due to a root-mean-square noise voltage
f_v = video-frequency bandwidth.

Then the slope of the side of a pulse is

$$\frac{H}{t_s} = \frac{N\sqrt{f_v}}{4}\Big/ b \qquad (37)$$

from which

$$b = \frac{N\sqrt{f_v}}{4}\,\frac{t_s}{H}. \qquad (38)$$

The audio signal recovered from the pulse wave, whether symmetrically or dissymmetrically modulated, is

$$\frac{CwH'}{\sqrt{2}}. \qquad (39)$$

In this expression, C is a proportional constant. H' is the new height of the pulse after the center section of the pulse has been selected out and the remainder rejected (as shown in Fig. 16). Amplitude due to the random displacement of one side of a pulse in dissymmetrical modulation is

$$CbH'. \qquad (40)$$

Hence the signal-to-noise ratio is

$$\frac{CwH'}{\sqrt{2}}\Big/ CbH' = \frac{2\sqrt{2}\,wH}{t_s N\sqrt{f_v}}. \qquad (41)$$

In symmetrical modulation, both sides of a pulse are subject to random displacement due to noise. The noise voltage given in (40) must therefore be multiplied by $\sqrt{2}$. The signal-to-noise ratio for symmetrical modulation is therefore

$$\frac{2wH}{t_s N\sqrt{f_v}}. \qquad (42)$$

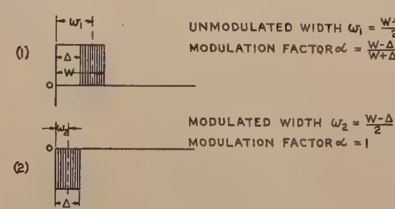


Fig. 17—Decomposition of a pulse time wave into width-modulated pulses (1) and (2) of the dissymmetrical type. Shaded lines intended to simulate an oscillogram.

4. *Pulses of Frequency Modulation*

The audio signal which may be recovered from a keyed frequency-modulated wave by means of a discriminator followed by a low-pass filter (Fig. 14) is proportional to the pulse width. However, the noise appearing in the audio output is proportional to the square root of the width. Hence

$$\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{P-F-M}} = \sqrt{r}\left(\frac{\text{signal}}{\text{noise}}\right)_{\text{standard F-M}}. \qquad (43)$$

Radio-Frequency Resistors as Uniform Transmission Lines*

D. ROGERS CROSBY†, ASSOCIATE, I.R.E., AND CAROL H. PENNYPACKER†

Summary—A theoretical study is made of the behavior of resistors, particularly the type where the resistance element is in the form of a film so there is negligible skin effect. When the electrical length of the resistor is a small fraction of a wavelength, it is shown that certain optimum proportions of the resistor exist in order best to terminate a radio-frequency transmission line.

Introduction

THE THEORETICAL and experimental behavior of resistors employed as transmission lines has been discussed frequently.[1,2] The material presented here is largely in the form of curves intended to give an easily grasped perspective of the subject. These curves are also suitable for design purposes. The dimensionless parameters used in plotting these curves are thought to be particularly convenient for engineerng use. The existence of certain optimum values of the parameters is shown.

This paper is an analysis of the classical transmission-line equations.We thus assume that the resistors employed as transmission lines are long compared to the diameter of the shields surrounding them, so that the current flowing in the short circuit at the far end has an electromagnetic field which is small compared to the total electromagnetic field surrounding the resistor. It is further assumed that the resistance per unit length is independent of frequency. This is substantially true in the film-type resistors used at radio frequencies, since the current penetration through the film is substantially complete for the usual resistance values. Particular attention has been given to the case where the resistor is intended to terminate or "*match*" a coaxial transmission line. A family of curves showing standing-wave ratio for various values of line resistance, as a function of frequency, has been plotted.

Consider the most common radio-frequency transmission line, made of copper. As the frequency of such a short-circuited line is raised from zero, the input resistance rises, reaching a maximum when the line is about a quarter wave long. For higher frequencies, the resistance oscillates with maxima slowly decreasing in amplitude. When the transmission-line conductor consists of a resistance that is not negligible compared to the characteristic impedance of the transmission line, the above impedance function is not obtained, since, as the frequency increases from zero, the input resistance may increase or decrease from the direct-current value. For frequencies a few times greater than the first resonant frequency, the input resistance may have a negligible amount of oscillation.

Consider a resistance employed as a transmission line and short-circuited at the far end (Fig. 1).

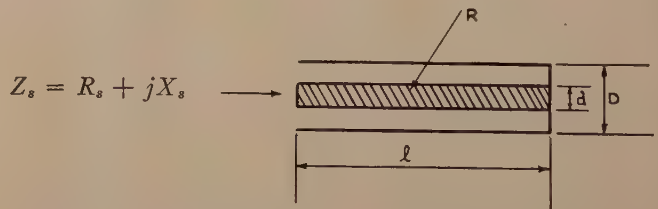


Fig. 1—Resistance employed as a transmission line. Far end is short-circuited.

l = length of line in inches
D = diameter of inside of outer conductor in inches
d = diameter of resistor in inches
Z_s = input impedance to line in ohms
Z_0 = characteristic impedance of line in ohms when R is zero
R = total value of series resistance of line in ohms
L = inductance of line in henries
C = capacitance of line in farads
f_{mc} = operating frequency in megacycles
λ = free-space wavelength in inches
$\omega = 2\pi f = 2\pi f_{mc} \times 10^6$

From the classical theory

$$Z_0 = 138 \log_{10} \frac{D}{d} \tag{1}$$

$$Z_0 = \sqrt{\frac{L}{C}} \tag{2}$$

$$C = \frac{l}{\lambda} \times \frac{1}{Z_0 \times f} \tag{3}$$

$$L = \frac{l}{\lambda} \times \frac{Z_0}{f} \tag{4}$$

* Decimal classification: R383×R144. Original manuscript received by the Institute, May 1, 1945; revised manuscript received, August 10, 1945.
† RCA Victor Division, Radio Corporation of America, Camden, N. J.
[1] G. H. Brown and J. W. Conklin, "Water-cooled resistors for ultra-high frequencies," *Electronics*, vol. 14, pp. 24–28; April, 1941.
[2] J. A. Fleming, "The Propagation of Electric Currents in Telephone and Telegraph Conductors," D. Van Nostrand Co., Inc., New York, N. Y., 1911 Chap. III, Eq. (51).

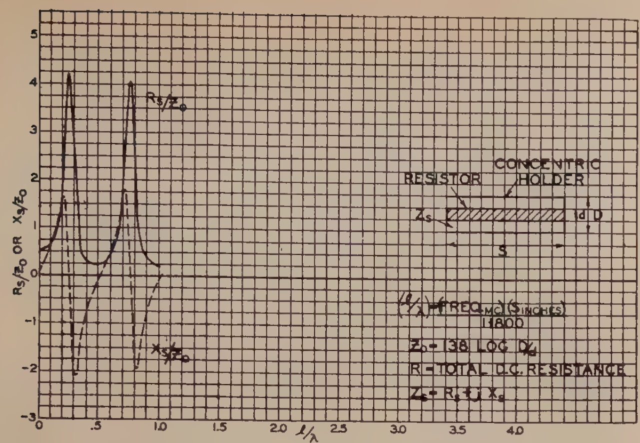


Fig. 2—Impedance of short-circuited transmission line. $R/Z_0 = 0.5$.

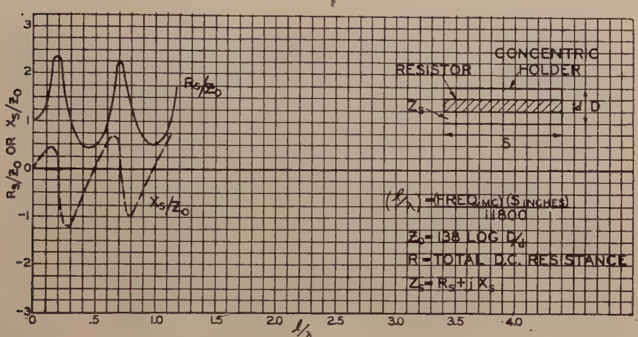


Fig. 3—Impedance of short-circuited transmission line. $R/Z_0 = 1.0$.

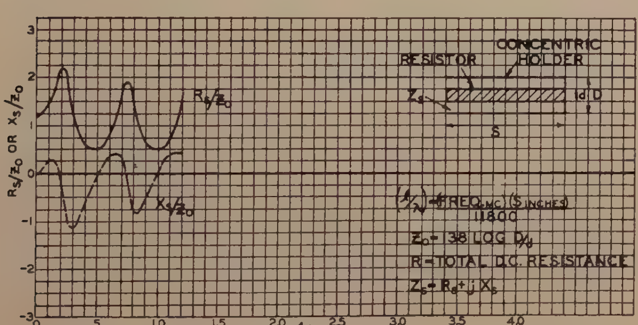


Fig. 4—Impedance of short-circuited transmission line. $R/Z_0 = 1.2$.

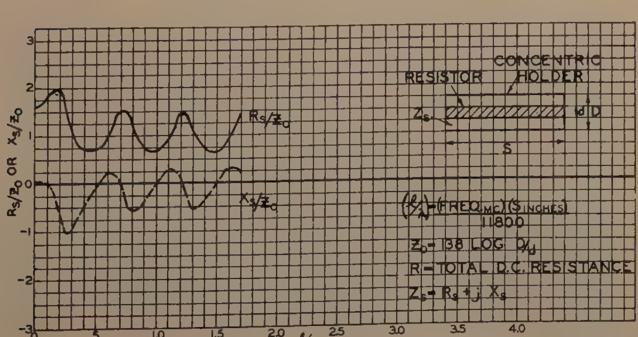


Fig. 5—Impedance of short-circuited transmission line. $R/Z_0 = 1.6$.

$$\frac{l}{\lambda} = \frac{l \times f_{mc}}{11{,}800} \tag{5}$$

$$Z_s = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \tanh \sqrt{(R + j\omega L)(G + j\omega C)}. \tag{6}$$

We consider only the case where G is negligibly small. From the above equations, we obtain

$$\frac{Z_s}{Z_0} = \sqrt{1 - j\frac{R}{Z_0}\frac{1}{2\pi(l/\lambda)}} \cdot \tanh \sqrt{-4\pi^2\left(\frac{l}{\lambda}\right)^2 + j2\pi\frac{l}{\lambda}\frac{R}{Z_0}}. \tag{7}$$

Two independent parameters are now involved, R/Z_0 and l/λ. The term R/Z_0 is independent of frequency, and is fixed by the proportion of the resistor to the jacket diameter. The term l/λ is proportional to frequency, so the plot of the impedance versus this term gives the frequency characteristic of the resistor.

The plot of (7) is given in Figs. 2, 3, 4, 5, 6, 7, 8, 9, and 10, and shows:

1. For values of R/Z_0 appreciably less than unity, both the resistance and reactance oscillate over a wide range of values.
2. For values of R/Z_0 much greater than unity, the oscillations are of minor amplitude.
3. For large values of R/Z_0, the reactance is always negative. For some short resistors, the reactance is positive, and for others it is negative.

Three special values of l/λ are of interest:

l/λ very small

l/λ very large

$l/\lambda = 1/4$.

Small Values of l/λ

Expanding (7) in powers of $2\pi l/\lambda$ we obtain

$$\begin{aligned} \frac{R_s}{Z_0} &= \frac{R}{Z_0} + (2\pi l/\lambda)^2\left(\left(\frac{R}{Z_0}\right)\frac{2}{3} - \frac{2}{15}\left(\frac{R}{Z_0}\right)^3\right) \\ &\quad + (2\pi l/\lambda)^4(\quad) + \cdots \\ \frac{X_s}{Z_0} &= (2\pi l/\lambda)\left(1 - \left(\frac{R}{Z_0}\right)^2\frac{1}{3}\right) \\ &\quad + (2\pi l/\lambda)^3(\quad) + \cdots . \end{aligned} \tag{8}$$

If the nth derivative of a function expressed in a power series is zero when the variable is zero, the coefficient of the nth term in the series must be zero.

Since in the expression for R_s/Z_0, the term involving (l/λ) to the first power is missing, we conclude that the slope of all the resistance curves will be zero at $l/\lambda = 0$. Figs. 2, 3, 4, 5, 6, 7, 8, 9, 10, and 11 illustrate this. For $R/Z_0 = \sqrt{5}$, the coefficient of the second term in the

expression for R_s/Z_0 vanishes. Thus for this value the curvature of the resistance characteristic will be zero at $l/\lambda=0$.

Since in the expression for X_s/Z_0, the term involving l/λ to the second power is missing, we conclude that the curvature of all the reactance curves (Fig. 12) will be zero at $l/\lambda=0$. For $R/Z_0=\sqrt{3}$, the coefficient of the first term in the expression for R_s/Z_0 vanishes. Thus, for this value, the slope of the reactance curve will be zero at $l/\lambda=0$.

When resistors are used for terminating transmission lines, the principal cause of standing waves is the presence of reactance. Thus the derived relation $R/Z_0=\sqrt{3}$ gives the widest frequency response for a fixed minimum standing-wave ratio. The RCA patent office called to our attention U. S. patent No. 2,273,547 filed in 1939 by Radinger of Berlin, Germany, which also gives the $\sqrt{3}$ ratio as being optimum. Radinger refers to German patent No. 618,678 filed in 1932 by Roosenstein. Roosenstein discloses the method of finding the optimum value. Due to an error in an expansion, he arrived at $\sqrt{2}$ instead of $\sqrt{3}$.

A plot has been made in Fig. 13 to show how the standing-wave ratio varies with frequency when a coaxial line is terminated with a resistor. The value of R is taken equal to the characteristic impedance of the line to be terminated. This insures that the line will be matched at low frequencies.

The value of R/Z_0 is determined by the ratio of the resistor diameter to the diameter of the transmission-line jacket in which the resistor is mounted. Thus the value of Z_0 is independent of the characteristic impedance of the line to be terminated. From Fig. 13 it can

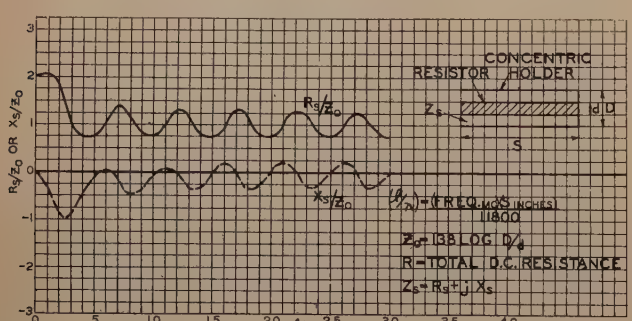


Fig. 6—Impedance of short-circuited transmission line. $R/Z_0=2.0$.

be seen that the standing-wave ratio nearest unity occurs for $R/Z_0=\sqrt{3}$.

Suppose we have a resistor 12 inches long and plan to use it up to 100 megacycles. Using (5) we obtain

$$l/\lambda = 0.1.$$

From the curves of Fig. 13, we see that if the jacket is chosen so $R/Z_0=\sqrt{3}$, the standing-wave ratio at 100 megacycles will be 0.9.

From this plot, we see that with the proper jacket, the

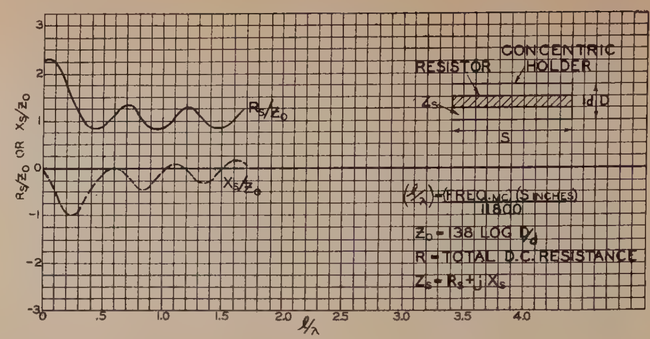


Fig. 7—Impedance of short-circuited transmission line. $R/Z_0=2.25$.

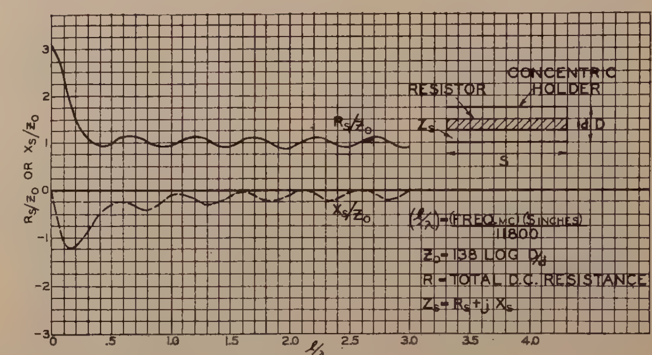


Fig. 8—Impedance of short-circuited transmission line. $R/Z_0=3.0$.

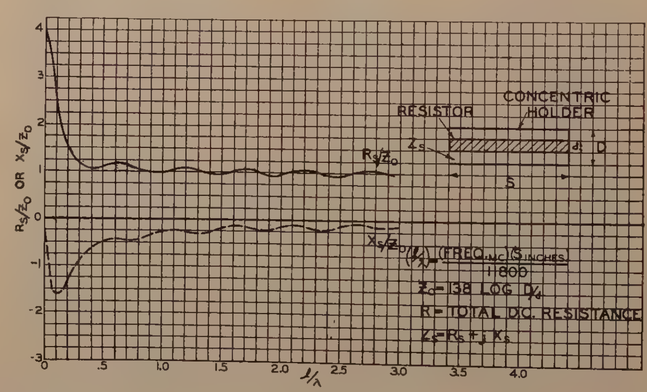


Fig. 9—Impedance of short-circuited transmission line. $R/Z_0=4.0$.

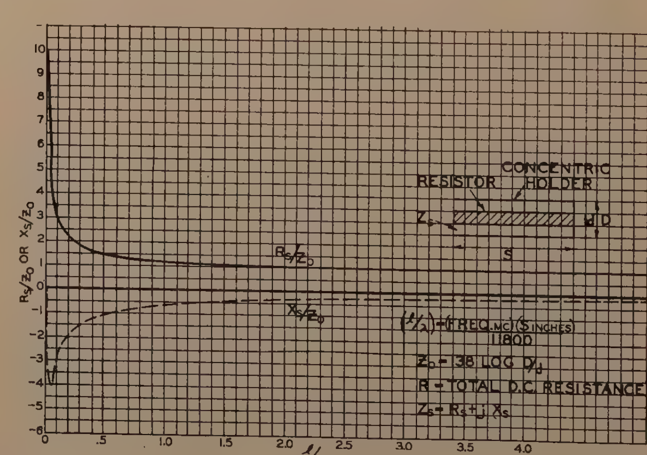


Fig. 10—Impedance of short-circuited transmission line. $R/Z_0=10.0$.

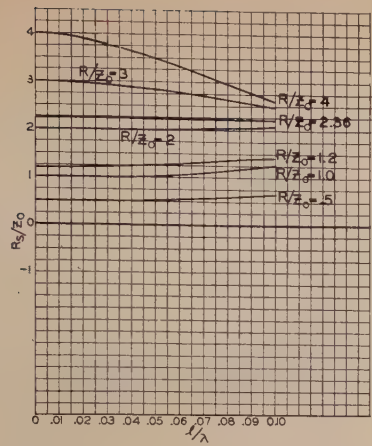


Fig. 11—Resistance of short-circuited transmission line. All slopes are zero at $l/\lambda=0$. For $R/Z_0=\sqrt{5}$, curvature is zero at $l/\lambda=0$.

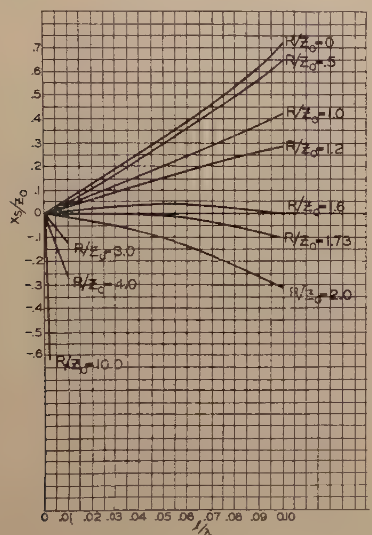


Fig. 12—Reactance of short-circuited transmission line. For $R/Z_0=\sqrt{3}$, slope is zero at $l/\lambda=0$.

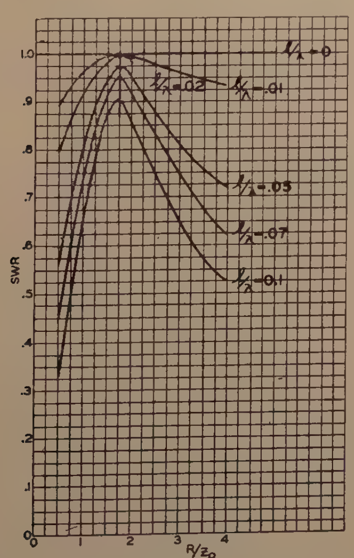


Fig. 13—Standing-wave ratio versus R/Z_0 for constant l/λ. Note as l/λ increases, the standing-wave ratio is minimum for $R/Z_0=\sqrt{3}$.

standing-wave ratio will be 0.97 for $l/\lambda=0.05$. Putting this value in (5)

$$0.05 = \frac{l \times f_{mc}}{11{,}800} \tag{9}$$

or approximately

$$f_{mc} = \frac{600}{l}.$$

This rounded constant 600 is convenient for quick design. Thus a resistor 1 inch long can have good characteristics up to 600 megacycles, and a resistor 12 inches long up to 50 megacycles.

The curves of Fig. 13 were computed using the formulas

$$\text{standing-wave ratio} = \frac{1-|K|}{1+|K|} \qquad K = \frac{1-\dfrac{Z_s}{R_0}}{1+\dfrac{Z_s}{R_0}} \tag{10}$$

where K *is the reflection coefficient.*

That the resistance may either increase or decrease as the frequency is raised from zero can also be shown from an equivalent lumped circuit of the resistor.[3,4] The equivalent circuit of the resistor is shown in Fig. 14, where $Z_0=\sqrt{L/C}$ and $Q=Z_0/R$.

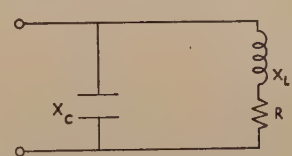


Fig. 14—Equivalent lumped circuit of resistor.

It can be shown that the reactance slope for the above circuit is zero at zero frequency for

$$\frac{R}{Z_0} = 1.$$

The condition at which the resistance begins to decrease as the frequency is raised is

$$\frac{R}{Z_0} = \sqrt{2}.$$

The corresponding two values from the transmission-line analysis are $\sqrt{3}$ and $\sqrt{5}$. Thus the lumped-circuit analysis gives only an approximate answer.

When resistors of the order of 10,000 ohms or higher

[3] D. B. Sinclair, "The type 663 resistor—a standard for use at high frequencies," *Gen. Rad. Exper.*, vol. 13, pp. 6–11; January, 1939.

[4] R. G. Anthes, "Behavior of resistors at radio frequencies," *Electronic Industries*, vol. 3, pp. 86–88; September, 1944.

are employed, it is not practical to mount them so that the Z_0 of the circuit is near the optimum value of

$$\frac{R}{Z_0} = \sqrt{3}.$$

When such resistors are mounted near a ground plane, R/Z_0 is usually several hundred. When such resistors are mounted well above a ground plane, the connecting leads to the resistor violate the assumption of this analysis.

Large Values of l/λ

To study the impedance characteristic for large values of l/λ, it is convenient to consider separately the two terms in our exact equation (7). The first term approaches a limit as l/λ becomes large.

$$\sqrt{1 - j\frac{R}{Z_0}\frac{1}{2\pi(l/\lambda)}} = \left[1 - j\left(\frac{R}{Z_0}\right)\frac{1}{4\pi(l/\lambda)} + \cdots\right]. \quad (11)$$

This term gives the normalized characteristic impedance of the line, and approaches unity in the limit as l/λ increases. Since the total resistance of the line is held constant, we should expect from physical reasoning that the impedance angle of the characteristic impedance would approach zero as the line length increases.

The second term of (7) does not approach a limit as l/λ increases, but oscillates between values dependent on R/Z_0. For l/λ large,

$$\tanh\sqrt{-4\pi^2(l/\lambda)^2 + j2\pi\frac{l}{\lambda}\frac{R}{Z_0}} = \frac{\tanh\dfrac{R}{2Z_0} + j\tan 2\pi\dfrac{l}{\lambda}}{1 + j\tanh\dfrac{R}{2Z_0}\tan 2\pi\dfrac{l}{\lambda}}. \quad (12)$$

As l/λ increases, the real part of this expression oscillates between $\tanh R/2Z_0$ and the reciprocal,

$$\frac{1}{\tanh\dfrac{R}{2Z_0}}.$$

The amplitude of the oscillation of the j part is much smaller, being

$$\frac{1}{2}\left[\tanh\frac{R}{2Z_0} - \frac{1}{\tanh\dfrac{R}{2Z_0}}\right].$$

This behavior is in contrast to the common case in which the frequency is held constant and the line length is increased. The total line resistance then increases indefinitely, and the input impedance of the line approaches a limit, which is the characteristic impedance of the line.

Consider the example plotted in Fig. 6.

$$\frac{R}{Z_0} = 2 \qquad \tanh\frac{R}{2Z_0} = 0.762.$$

The normalized resistance oscillates for large l/λ between 0.76 and 1.31 while the reactance oscillates between ± 0.28. These values are obtained by substituting in the above expressions. The plot in Fig. 6 shows how the oscillation approaches a finite limit for its amplitude.

The oscillations of the plot of Fig. 10 are negligible since $R/Z_0 = 10$ and $\tanh R/2Z_0 = 0.9999$.

Discussion of $l/\lambda = 1/4$

For small value of R/Z_0, the resistance maximizes near $l/\lambda = 1/4$.

The value of this maximum resistance is approximately

$$\left(\frac{R}{Z_0}\right)_{\max} = \frac{2Z_0}{R}.$$

In Fig. 2, $R/Z_0 = 0.5$ and the first maximum is approximately 4, as the plot shows. This situation has been discussed by Terman.[5]

[5] F. E. Terman, "Resonant lines in radio circuits," *Elec. Eng.*, vol. 53, pp. 1046–1053; July, 1934.

An Analysis of Three Self-Balancing Phase Inverters*

MYRON S. WHEELER†, ASSOCIATE, I.R.E.

Summary—**A self-balancing phase inverter is a circuit converting one driving voltage to two output voltages of opposite phase but of essentially equal magnitude by an inherent characteristic of the device and not by virtue of any critical adjustment. The algebraic solution of three self-balancing phase inverters is given, assuming all circuit elements are linear. Included in the solution are the conditions for self-balance, the balance ratio, and the voltage gain. From this information, the type of inverter for a particular service may be selected and designed.**

Introduction

THE DESIGN of audio and video amplifiers frequently requires the conversion from a single-ended to a double-ended (or push-pull) channel. Two familiar examples of this type of amplifier are the driving of cathode-ray-tube plates in push-pull from an amplifier that must be, for convenience, single ended at its input; and the driving of audio power amplifiers in push-pull where it is again convenient to use single-ended voltage-amplifier stages and input.

This conversion requires a network with two output voltages, equal in magnitude but opposite in phase, and proportional to its driving voltage. While a transformer fulfills these requirements, it is frequently more economical and more uniform in frequency response to use a tube to invert the phase in a circuit that automatically equalizes the two output signals.[1,2]

Three self-balancing phase-inverting circuits will be analyzed: the common-plate-impedance inverter, the common-cathode-impedance inverter, and the cathode- and plate-loaded inverter. The only assumption made in the analysis is that all circuit elements (and in particular, the tube parameters) are linear. This assumption is well justified, as this type of amplifier is generally designed for linear operation. And, although the solutions are given for such a frequency range that the load impedances are resistive, the solution could be extended to any frequency by the substitution of their complex impedances. The circuits in Figs. 1, 2, 3, 4, 5, and 6 are all equivalent signal-voltage circuits, as there is no loss in generality by omitting the direct-current components.

Common-Plate-Impedance Self-Balancing Inverter

Referring to Figs. 1 and 2, the following tube parameters are defined as:

$$r_{p1} = \frac{\partial E_{p1}}{\partial i_1} \qquad \mu_1 = \frac{\partial E_{p1}}{\partial e_g}$$

$$r_{p_2} = \frac{\partial E_{p_2}}{\partial i_2} \qquad \mu_2 = \frac{\partial E_{p_2}}{\partial e_0}.$$

Then, adding the voltages around three closed circuits,

$$\mu_1 e_g - i_1 r_{p1} - i_1 r_1 - (i_1 - i_2)R_0 = 0$$

$$\mu_2 e_0 - i_2 r_{p_2} - i_2 r_2 + (i_1 - i_2)R_0 = 0$$

$$e_0 - (i_1 - i_2)R_0 = 0$$

and as

$$E_2 = i_2 r_2 - e_0$$

and

$$E_1 = i_1 r_1 + e_0.$$

Solving for E_2/E_1

$$\frac{E_2}{E_1} = \frac{\mu_2 R_0 r_2 - r_{p_2} R_0}{\mu_2 R_0 r_1 + r_1 r_2 + r_1 r_{p_2} + R_0 r_1 + R_0 r_2 + R_0 r_{p_2}}. \qquad (1)$$

In (1) the terms $\mu_2 R_0 r_2$ and $\mu_2 R_0 r_1$ are much greater than the rest, and it can be seen that, if the others were

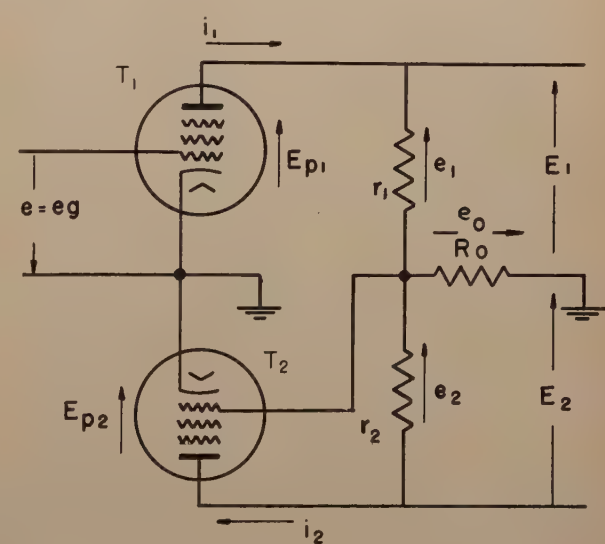


Fig. 1—Equivalent signal-voltage circuit of common-plate-impedance self-balancing inverter.

negligible compared with these two, the condition for balance (equal output voltages) would be

$$\mu_2 R_0 r_2 = \mu_2 R_0 r_1 \quad \text{or} \quad r_2 = r_1.$$

The circuit could be brought to balance, however, in

* Decimal classification: R139. Original manuscript received by the Institute, July 6, 1945; revised manuscript received, October 2, 1945.
† Westinghouse Electric Corporation, Bloomfield, N. J.
[1] J. G. Brainerd, "Ultra-High-Frequency Techniques," D. Van Nostrand Co., Inc., New York, N. Y., 1942, pp. 100–101.
[2] M.I.T. Staff, "Applied Electronics," John Wiley and Sons, New York, N. Y., 1943, pp. 489–490.

any event by adjusting r_1 and r_2 so that

$$\frac{E_2}{E_1} = 1$$

but we are more interested in the degree of unbalance resulting when $r_2 = r_1$, which we shall call r; then

$$\frac{E_2}{E_1} = \frac{\mu_2 R_0 r - r_{p_2} R_0}{\mu_2 R_0 r + r^2 + r r_{p_2} + 2R_0 r + R_0 r_{p_2}}. \tag{2}$$

In order to minimize the unbalance, R_0 should be selected as great as possible, as it can be seen from (2) that

$$\frac{E_2}{E_1} \rightarrow \frac{\mu_2 r - r_{p_2}}{\mu_2 r + 2r + r_{p_2}} \tag{3}$$
$$\lim R_0 \rightarrow \infty$$

which is the closest E_2/E_1 comes to 1 as R_0 varies from 0 to ∞. Practical limits, however, set the maximum

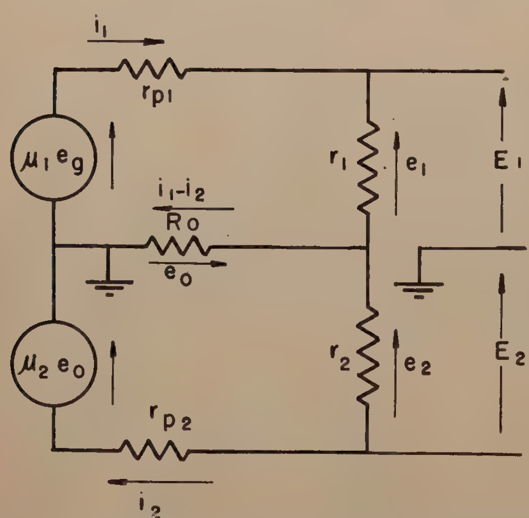


Fig. 2—Equivalent linear-tube-parameter circuit of common-plate-impedance self-balancing inverter.

value of R_0 near r. As a basis of comparison of this type of phase inverter with some others, let $R_0 = r$. Then

$$\frac{E_2}{E_1} = \frac{\mu_2 r - r_{p_2}}{\mu_2 r + 3r + 2r_{p_2}}. \tag{4}$$

To simplify this expression further, if we define the gain of the inverter tube T_2 as N, and omit intermediate algebraic steps

$$N = \frac{E_2}{e_0} = \frac{\mu_2 r_2 - r_{p_2}}{r_{p_2} + r_2} \tag{5}$$

then with $r_1 = r_2 = r$ as above from (2)

$$\frac{E_2}{E_1} = \frac{NR_0}{r + R_0(N+2)} \tag{6}$$

and with $R_0 = r$ as in (4)

$$\frac{E_2}{E_1} = \frac{N}{N+3} \text{ (exactly)} \tag{7}$$

where N, the gain of the inverter tube T_2, is defined exactly by (5). It is essentially, however, the gain of T_2 as a normal plate-loaded voltage amplifier.

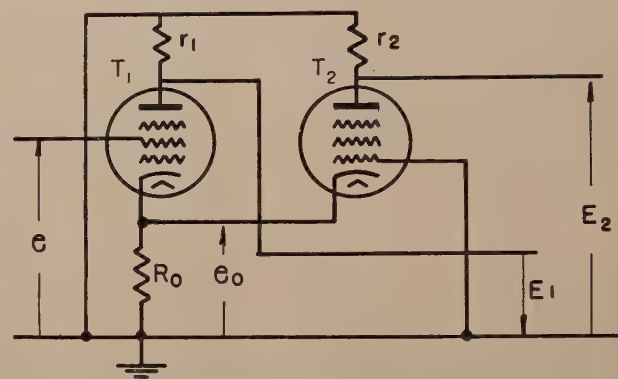


Fig. 3—Equivalent signal-voltage circuit of common-cathode-impedance self-balancing inverter.

That is

$$N \cong \text{normal plate-loaded gain} = \frac{\mu_2 r_2}{r_{p_2} + r_2}$$

as can be seen from (5), because generally

$$\mu_2 r_2 \gg r_{p_2}.$$

Let us stop to interpret these equations. We find from (7) that the balance of the inverter depends upon the gain of T_2, provided R_0 is sufficiently great as shown in (3), where the gain of T_2 is defined by (5). The balance also requires that the plate-load resistors of the two stages be equal as seen in (1), but the two tubes need not be the same. The tube T_1, then, acting only as a voltage amplifier has a gain

$$\frac{E_1}{e} = \frac{\mu_1\mu_2 R_0 r_1 + \mu_1(r_1 r_2 + r_1 r_{p_2} + r_1 R_0 + R_0 r_2 + R_0 r_{p_2})}{(r_1 + r_{p_1})(r_2 + r_{p_2}) + R_0(r_1 + r_{p_1})(1+\mu_2) + R_0(r_2 + r_{p_2})}. \tag{8}$$

The only purpose of writing (8) was to show approximately that the gain of T_1 is the same as a normal plate-loaded amplifier with a load resistance of r_1, if

$$\gg 1$$

for then

$$\frac{E_1}{e_g} \cong \frac{\mu_1\mu_2 R_0 r_1}{(\mu_2 + 1)(r_1 + r_{p_1})R_0} \cong \frac{\mu_1 r_1}{(r_1 + r_{p_1})}. \tag{9}$$

Common-Cathode-Impedance Self-Balancing Inverter

Using the same symbols as in the previous analysis, it can be seen from Fig. 3 that

$$e_{g1} = e - (i_1 + i_2)R_0$$
$$e_{g2} = (i_1 - i_2)R_0.$$

Adding the voltages around two closed paths in Fig. 4

$$(i_1 - i_2)R_0 - \mu_1[e - (i_1 - i_2)R_0] + i_1(r_1 + r_{p1}) = 0$$
$$(i_1 - i_2)R_0 + \mu_2(i_1 - i_2)R_0 - i_2(r_{p2} + r_2) = 0.$$

Then solving for E_2/E_1

$$\frac{E_2}{E_1} = \frac{R_0(\mu_2 + 1)r_2}{R_0(\mu_2 + 1)r_1 + (r_{p2} + r_2)r_1}. \quad (10)$$

As in the previous solution, let

$$r_1 = r_2 = r.$$

R_0 should again be as great as practical limitations permit to obtain the best balance. In general this is near the value of r, and to compare this inverter with the previous type, let

$$R_0 = r_1 = r_2 = r.$$

Then

$$\frac{E_2}{E_1} = \frac{\mu_2 r + r}{\mu_2 r + 2r + r_{p2}}. \quad (11)$$

This expression may also be simplified in the same manner as the previous case.

The common-cathode-impedance phase inverter, it has been seen, is quite analogous to the common-plate-

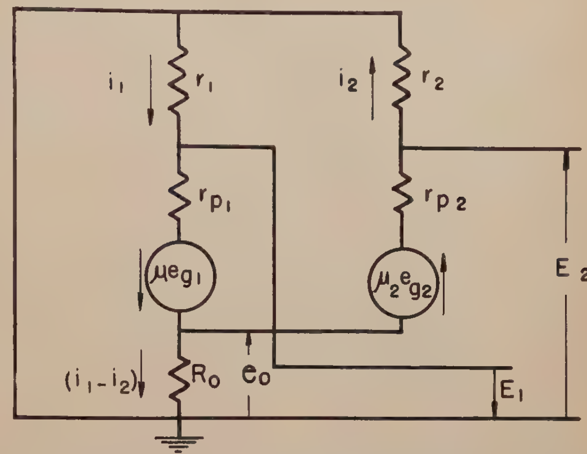


Fig. 4—Equivalent linear-tube-parameter circuit of common-cathode-impedance self-balancing inverter.

impedance phase inverter. Good balance depends upon equal load resistors, high gain in T_2, and a common cathode resistor as large as practicable. The degree of balance is a little better than the common-plate-impedance type, but the voltage amplification of T_1 is found to be about half, for

$$\frac{E_1}{e} = \frac{\mu_1(1 + \mu_2)R_0 r_1 + \mu_1(r_{p2} + r_2)r_1}{(\mu_2 + 1)(r_1 + r_{p1})R_0 + (\mu_1 + 1)(r_2 + r_{p2})R_0 + (r_1 + r_{p1})(r_2 + r_{p2})}. \quad (15)$$

Define the gain of T_2 as

$$N = \frac{E_2}{e_0} = \frac{(\mu_2 + 1)r_2}{r_{p2} + r_2}. \quad (12)$$

Then

$$\frac{E_2}{E_1} = \frac{R_0 N}{R_0 N + r} \quad (13)$$

and with

$$R_0 = r$$

$$\frac{E_2}{E_1} = \frac{N}{N + 1} \text{ (exactly)} \quad (14)$$

where N is again essentially the gain of T_2 as a normal plate-loaded voltage amplifier.

That is,

$$N \cong \text{normal plate-loaded gain} = \frac{\mu_2 r_2}{r_{p2} + r_2}$$

as it can be seen from (10), because generally

$$\mu_2 \gg 1.$$

To simplify the equation consider

$$r_1 + r_{p2} = r_2 + r_{p2}$$

and

$$\mu_1 = \mu_2$$

where

$$\mu_2 \gg 1.$$

Then

$$\frac{E_1}{e} = \frac{\mu_1 r_1}{2(r_1 + r_{p1})}. \quad (16)$$

Again the only purpose of writing (15) was to show approximately the gain of T_1, which is roughly one half of the gain of the normal plate-loaded amplifier with a load resistance of r_1.

Cathode- and Plate-Loaded Inverter

Again using the same tube parameters, it can be seen from Fig. 5 that

$$e_g = e - E_2$$

and from Fig. 6

$$i_p = \frac{\mu e_g}{r_p + r_1 + r_2}.$$

Then solving for E_2/E_1

$$\frac{E_2}{E_1} = \frac{r_2}{r_1}$$

and perfect balance is obtained when

$$r_2 = r_1.$$

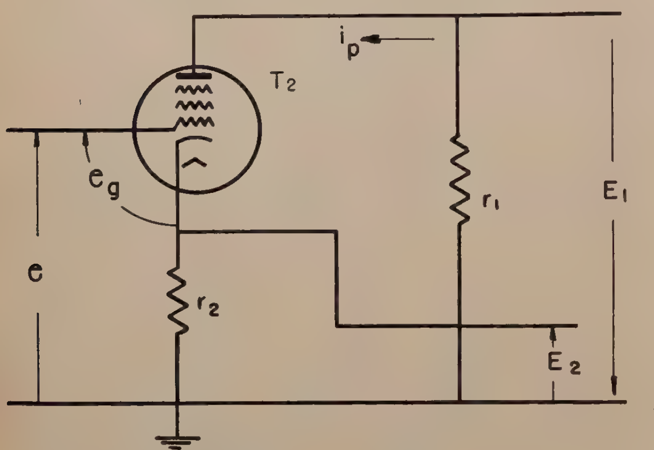


Fig. 5—Equivalent signal-voltage circuit of cathode- and plate-loaded inverter.

Define

$$N = \frac{E_2}{e}$$

and making

$$r_2 = r_1 = r$$

$$N = \frac{\mu r}{r_p + 2r + \mu r}. \quad (17)$$

The gain is essentially unity as seen from (17) since generally

$$\mu r >> r_p + 2r.$$

The conditions for balance are less rigid in this type of inverter, the only requirement being that the load resistors be equal. To compare the voltage gain of this single-tube circuit with the previous types, let us consider a voltage amplifier before the inverter as part of the voltage gain. This tube would have a gain

$$\frac{E}{e} = \frac{\mu r}{r_p + r}. \quad (18)$$

The inverter, having an approximate gain of unity per phase, will give an over-all gain, to a first approximation, equal to the gain of the first tube. This is also approximately the gain obtainable with the common-plate-loaded inverter, the exact ratios of the gains being easily obtainable but of little interest.

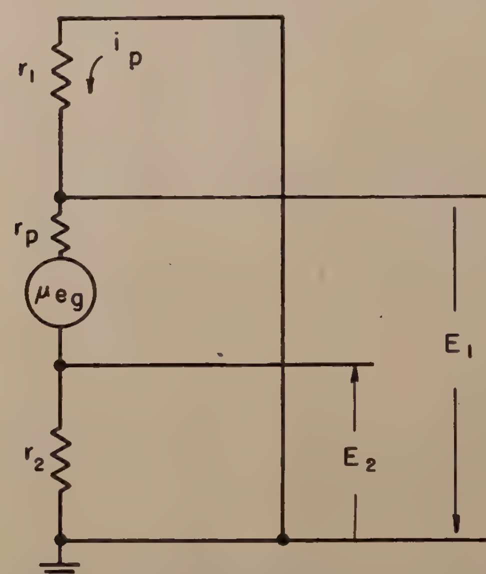


Fig. 6—Equivalent linear-tube-parameter circuit of cathode- and plate-loaded inverter.

Table I summarizes this data so that the three phase inverters may be compared.

TABLE I

Type	Balance Depends Upon	Exact Degree of Balance	Exact Inverter Gain	Approximate Inverter Gain	Approximate Relative Over-All Voltage Gain
Common Plate Load	$r_1 = r_2 = r$ $R_0 \approx r$ $N >> 3$	$\frac{N}{N+3}$	$N = \frac{\mu_2 r_2 - r_{p_2}}{r_{p_2} + r_2}$	$N \cong \frac{\mu_2 r_2}{r_{p_2} + r_2}$	1
Common Cathode Load	$r_1 = r_2 = r$ $R_0 \approx r$ $N >> 1$	$\frac{N}{N+1}$	$N = \frac{(\mu_2 + 1) r_2}{r_{p_2} + r_2}$	$N \cong \frac{\mu_2 r_2}{r_{p_2} + r}$	$\frac{1}{2}$
Plate-Loaded Cathode-Loaded	$r_1 = r_2 = r$	1	$N = \frac{\mu r}{r_p + 2r + \mu r}$	$N \cong 1$	1

Pulse Response of Thyratron Grid-Control Circuits*

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Summary—For applications in which a thyratron must function as an accurately controlling device, rather precise grid control is required. Frequently, the desired precision of firing is obtained by applying peaked voltage pulses to the grid of the thyratron. Satisfactory operation of the thyratron and its associated grid-control circuit usually requires an external grid-cathode capacitance and a current-limiting grid resistance. For a given pulse shape supplied by the grid signal generator, the grid resistance and capacitance can alter materially the wave form of the voltage which appears on the grid. This paper presents, in the form of curves, an analysis of the influence of the grid-circuit components upon the grid response for several commonly used grid signal pulses. Advantages of peaked wave-form grid control are discussed, and an example is worked out to illustrate the use of the curves in evaluating the grid response to a typical signal pulse.

I. Introduction

THE LAST FEW years have seen a gradual increase in the use of peaked wave-form grid signals for thyratron control. Several of the advantages of peaked-grid-signal control can be seen by reference to Figs. 1 and 2. The various grid-control curves shown in Fig. 1 may differ from each other for several reasons: (1) the control characteristics corresponding to different condensed-mercury temperatures for a mercury-vapor tube; (2) random variation in control characteristic from tube to tube of the same type; (3) control curves corresponding to different types of tubes, interchangeable, except for grid characteristics. In Fig. 2 the control curves of Fig. 1 have been trans-

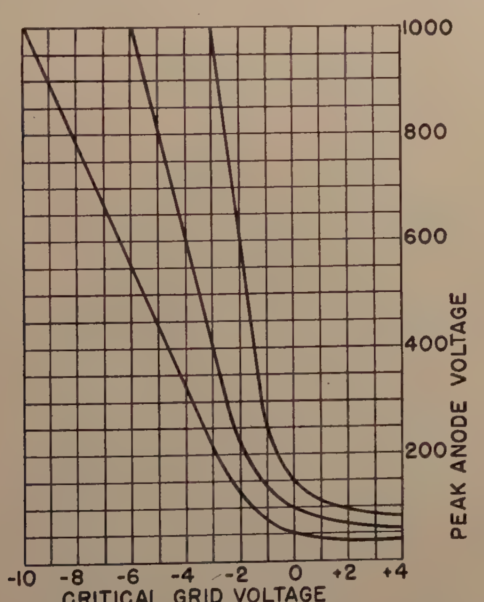

Fig. 1—Typical grid-control curves for a thyratron.

lated in the conventional manner for a sinusoidal anode voltage applied to the tube. Thus, for the positive half cycle of anode voltage shown in Fig. 2, the corresponding time variation of the critical grid potential is determined by the various curves of Fig. 1.

Now, by applying a grid signal with a very high rate of rise we can "sweep" through the various grid-control

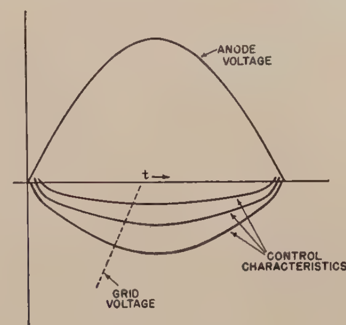

Fig. 2—Grid-control curves of Fig. 1 for a sinusoidal anode voltage.

characteristics very rapidly. The intersections of this steep-wave-front signal with the characteristic curves determine the instants of time at which the thyratron might fire, depending, of course, upon the particular grid characteristic the tube might have. By increasing the grid-signal rate of rise, it is possible to confine these intersections, or firing points, to a very narrow time interval. In this manner very precise firing can be obtained in spite of considerable variations in control characteristic.

To obtain a comparable rate of rise (and hence the same precision of firing) by means of an ordinary alternating-current grid signal would require an extremely large amplitude, which in most cases would be considerably in excess of the maximum allowable grid voltage specified by the tube manufacturer. Furthermore, over a portion of the range of phase control, an alternating-current signal will drive the grid positive during the inverse cycle of anode voltage. Such a condition of positive grid potential while the anode is negative is thought to increase materially the probability of an arc back; that is, failure of the tube to stand inverse voltage. There is also some evidence to indicate that this condition may increase the rate of "cleanup" in tubes filled with inert gasses. These difficulties are easily avoided by the use of a narrow peaked wave for the grid signal.

A general discussion of peaked-wave-form grid control, including several other advantages such as short ionization and deionization times, has been given by Morack.[1]

A considerable amount of material on pulse-forming circuits may be found in the literature.[2,3] Several of these have been adapted to thyratron-control applications by incorporating with them a phase-shifting

* Decimal classification: 621.375.1. Original manuscript received by the Institute, May 14, 1945.
† Westinghouse Electric Corporation, Bloomfield, N. J.

[1] M. M. Morack, "Voltage impulses for thyratron grid control," *Gen. Elec. Rev.*, vol. 37, pp. 288–295; June, 1934.
[2] O. Kiltie, "Transformers with peaked waves," *Elec. Eng.*, vol. 51, pp. 802–804; November, 1932.
[3] J. G. Brainerd, Glenn Koehler, Herbert J. Reich, and L. F. Woodruff, "Ultra-High-Frequency Techniques," D. Van Nostrand Co., Inc. New York, N. Y., 1942, Chapter 4.

network, enabling the peak of the grid pulse to be shifted in time with respect to the thyratron's anode voltage.

II. Grid-Circuit Response to Peaked-Wave-Form Grid Signals

The grid-control circuits used in the majority of thyratron applications can be simplified for this analysis into the circuit of Fig. 3. The bias battery prevents firing of the thyratron until a definite grid signal is applied. In practical circuits, the bias battery is usually

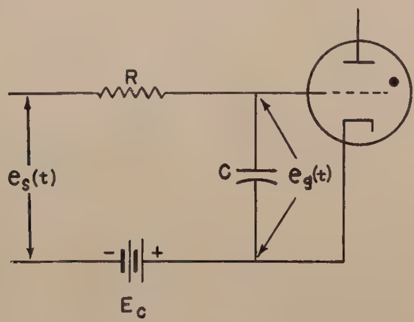

Fig. 3—Typical grid-control circuit for a thyratron.

replaced by a dry-type rectifier, tube rectifier, or an alternating-current voltage. The grid resistance limits the current which may be drawn from the grid-voltage supply. The grid capacitance C is the equivalent capacitance of the grid-cathode tube capacitance, combined with the external capacitance. This external capacitance is usually added in conjunction with the grid resistance to minimize undesirable fluctuations in grid potential resulting from disturbances coming in from the anode through the electrostatic coupling between grid and anode, or disturbances coming in through the grid circuit itself.

For most applications, the grid resistance R varies from 1000 ohms to several megohms and the capacitance C ranges from a few micromicrofarads to several hundredths of a microfarad. The grid-cathode capacitance of the thyratron is almost always very much less than the capacitance added externally, so it can usually be neglected. If preconduction grid currents are ignored, the influence of the tube on its grid potential is negligible until the initiation of the discharge between cathode and anode. (This assumption will be discussed later.)

However, the grid resistance and capacitance can affect very materially the wave form and magnitude of the potential which appears on the grid for a given wave form from the grid signal generator. The alteration in the grid-potential wave form is, of course, more serious the steeper the wave front of the signal applied to the circuit. Since this distortion in wave form can be very appreciable, it seems of interest to examine the response of the grid circuit for several types of signal impulses commonly used. For simplicity it will be assumed that the internal impedance of the signal generator is small compared to the impedance of the grid circuit, so that the generator's wave form is not influenced appreciably by the loading imposed by R and C. This assumption is a fairly reasonable one for the signal generators usually employed in this type of application and for the ranges of R and C required for satisfactory tube operation.

The response to three types of pulses has been analyzed. These are the triangular, square, and exponential pulses shown in Fig. 4. Only the results are presented in the main body of the paper. The essential details of the analysis for each type of pulse may be

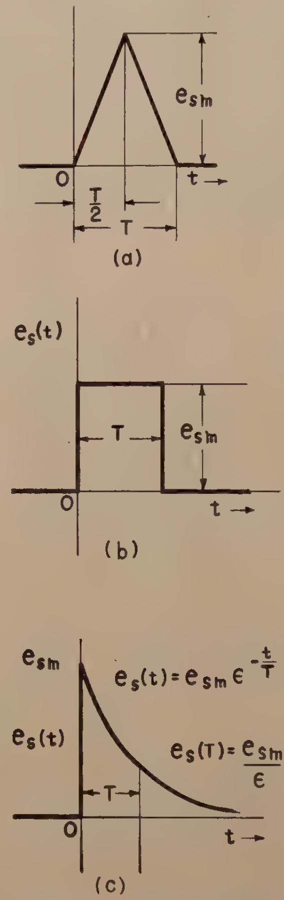

Fig. 4—Grid-signal pulses.

found in the mathematical appendix. For generality the analytical results are presented in the form of curves. These curves, showing the response characteristics of the grid circuit, involve dimensionless quantities representing the parameters of the grid circuit and the signal pulses.

a. Response to a triangular pulse: The analysis of the response to a triangular pulse of duration T is quite simple if the pulse is regarded as being formed by the addition of three functions as shown in (a) in Fig. 5. Resolving the pulse into these three components leads to three expressions for the response, each valid for the respective time intervals $0 \leqq t \leqq T/2$, $T/2 \leqq t \leqq T$, $t \geqq T$.

Nomenclature

t = time
$e_g(t)$ = instantaneous voltage between grid and cathode
e_{gm} = maximum value of the grid-to-cathode voltage
$e_s(t)$ = instantaneous electromotive force of the grid-signal generator
e_{sm} = maximum value of the signal pulse
$i(t)$ = instantaneous current in the grid circuit
E_c = grid-bias-supply voltage
T = time of duration of the grid-signal pulse
R = grid resistance
C = grid-cathode capacitance
$\beta = RC/T$
Δt = time interval between instant of maximum signal-pulse voltage and the instant of maximum grid-to-cathode voltage

τ = time required for the grid-potential wave form to pass through two critical values of grid potential
ln = logarithm to the base ϵ
p = parameter of the Laplace transform
$E_g(p)$ = Laplace transform of $e_g(t)$
$E_s(p)$ = Laplace transform of $e_s(t)$
$I(p)$ = Laplace transform of $i(t)$

For no initial charge on the capacitance C, the ratio of the instantaneous grid potential to the maximum value of the signal pulse $(e_g(t)/e_{sm})$ is found to be

$$\frac{e_g(t)}{e_{sm}} = 2\left\{\frac{t}{T} - \beta(1 - \epsilon^{-(1/2)(t/T)})\right\} \quad \text{for } 0 \leqq t \leqq T/2$$

$$= 2\left\{\left(1 - \frac{t}{T}\right) + \beta[1 + \epsilon^{-(1/\beta)(t/T)}(1 - 2\epsilon^{1/2\beta})]\right\} \quad \text{for } T/2 \leqq t \leqq T$$

$$= 2\beta\epsilon^{-(1/\beta)(t/T)}\{1 - 2\epsilon^{1/2\beta} + \epsilon^{1/\beta}\} \quad \text{for } t \geqq T \tag{1}$$

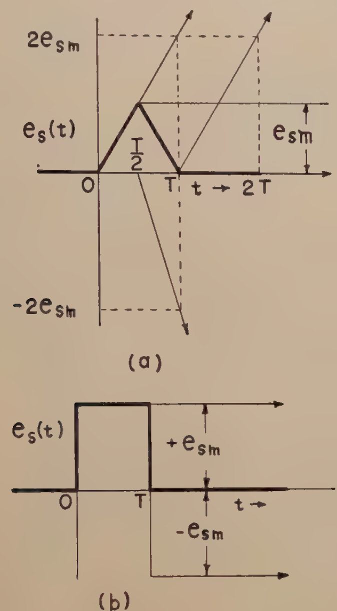

Fig. 5—Components of the triangular and square pulses.

where $\beta = RC/T$. In Fig. 6 the response of (1) is shown as a family of curves in terms of the parameter β and the variable t/T. Examination of the time derivatives of (1) shows that the maximum value of the ratio $e_g(t)/e_{sm}$ always occurs during the interval $T/2 \leqq t \leqq T$ at an instant t_m obtainable from

$$t_m/T = \beta ln(2\epsilon^{1/2\beta} - 1). \tag{2}$$

The ime interval Δt between the peak of the signal pulse and the peak grid voltage can then be expressed in terms of a ratio to the pulse width T as

$$\Delta t/T = \frac{t_m - T/2}{T} = \beta ln(2\epsilon^{1/2\beta} - 1) - 1/2. \tag{3}$$

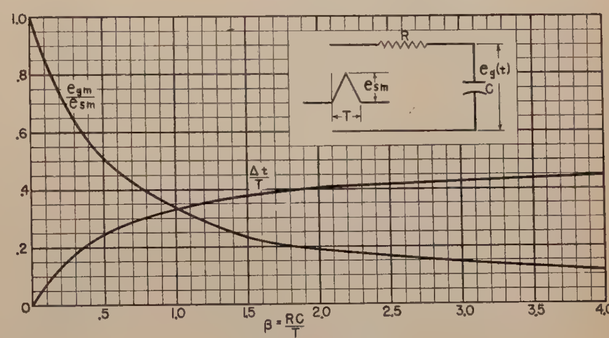

Fig. 7—The ratios (e_{gm}/e_{sm}) and $(\Delta t/T)$ as functions of $\beta = RC/T$ for a triangular pulse.

This ratio $\Delta t/T$ is shown as a function of β in Fig. 7. Note that the curve approaches the value 0.5 as a limit for very large values of β as t_m must occur in the range $T/2 \leqq t_m \leqq T$. Substitution of the value t_m from (2) into (1) gives as the maximum value of $(e_g(t)/e_{sm})$

$$e_{gm}/e_{sm} = 2[1 - \beta ln(2\epsilon^{1/2\beta} - 1)], \tag{4}$$

and from (3) this ratio is also

$$e_{gm}/e_{sm} = 1 - 2\left(\frac{\Delta t}{T}\right). \tag{5}$$

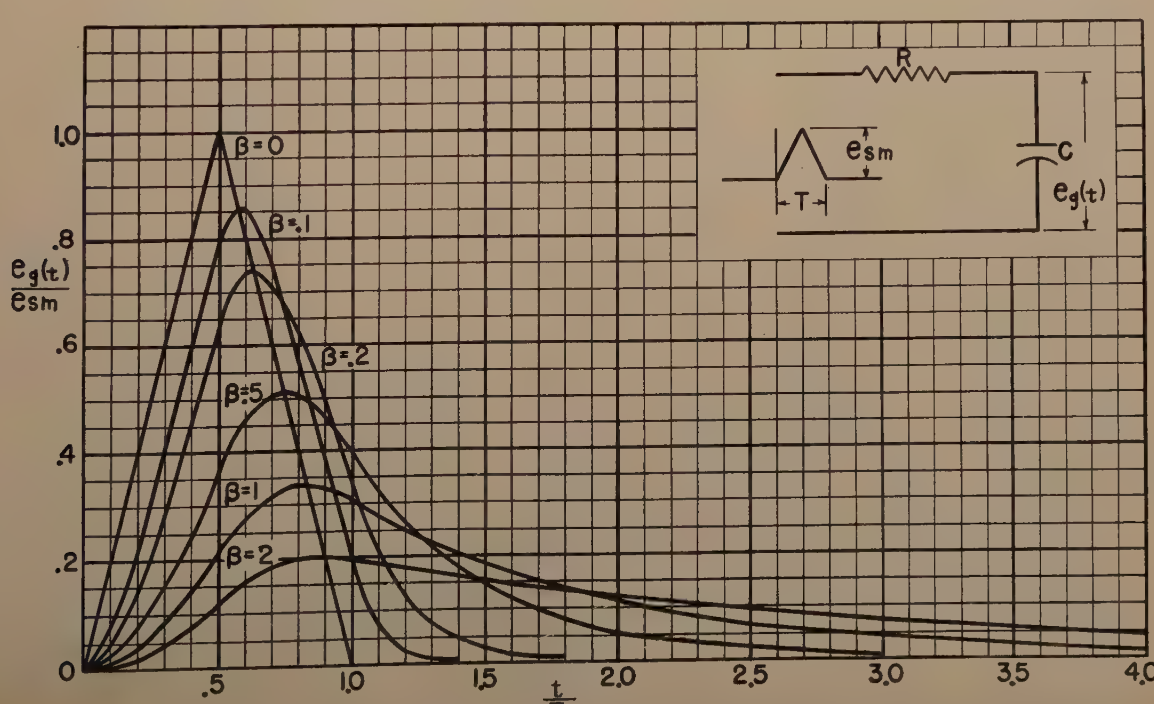

Fig. 6—The ratio $(e_g(t)/e_{sm})$ as a function of (t/T) for a triangular pulse with $\beta = RC/T$ as a parameter.

This ratio is shown as a function of β in Fig. 7 and in terms of $(\Delta t/T)$ in Fig. 8.

b. Response to a square pulse: The square pulse of Fig. 4 (b) is often regarded as being formed by two step functions as shown by (b) in Fig. 5. Correspondingly, the response is found to consist of two expressions, each valid over different time intervals. For the grid capacitor initially uncharged, the ratio of the grid voltage to the height of the square pulse is

$$\begin{aligned} e_g(t)/e_{sm} &= 1 - \epsilon^{-(1/\beta)(t/T)} \quad \text{for } 0 \leqq t \leqq T \\ &= \epsilon^{-(1/\beta)(t/T)}(\epsilon^{1/\beta} - 1) \quad \text{for } t \geqq T. \end{aligned} \tag{6}$$

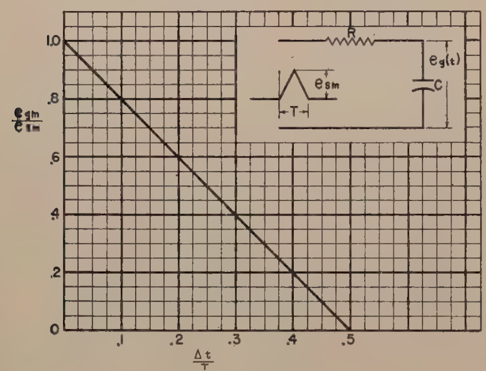

Fig. 8—The ratio (e_{gm}/e_{sm}) as a function of $(\Delta t/T)$ for a triangular pulse.

This function is presented in Fig. 9 as a family of curves with β as the parameter and t/T as the independent variable where, as before, $\beta = RC/T$. By examination it is evident that the grid voltage is a maximum at the end of the pulse, so

$$t_m = T. \tag{7}$$

The ratio of the maximum grid voltage to the height of the square pulse is, by (6) and (7)

$$e_{gm}/e_{sm} = 1 - \epsilon^{-1/\beta}. \tag{8}$$

This ratio is shown as a function of β in Fig. 10.

c. Response to an exponential pulse: The exponential pulse of Fig. 4 (c) has a maximum value of e_{sm} at $t=0$ and a time constant T. In a sense, T may be regarded as the pulse width, in that it represents the width of the pulse at the instant for which $e_s(t) = e_{sm}/\epsilon$. The response of the grid circuit to an exponential pulse falls into two cases: the general case $RC \neq T$, and the special case $RC = T$. For the grid capacitor initially uncharged, the

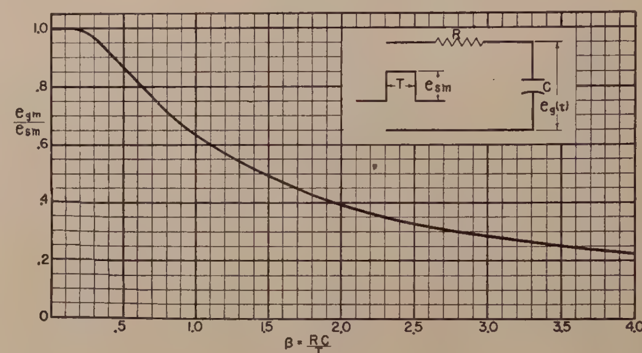

Fig. 10—The ratio (e_{gm}/e_{sm}) as a function of $\beta = RC/T$ for a square pulse.

ratio of the grid-capacitor voltage to the maximum value of the exponential pulse is, in terms of $\beta = RC/T$,

$$\begin{aligned} e_g(t)/e_{sm} &= \epsilon^{-t/T} - \epsilon^{-(1/\beta)(t/T)} \quad \text{for } RC \neq T \\ &= \frac{t}{T}\epsilon^{-t/T} \quad \text{for } RC = T. \end{aligned} \tag{9}$$

This ratio $(e_g(t)/e_{sm})$ is shown as a function of β and t/T in Fig. 11. The time derivatives of (9) show that

$$\frac{t_m}{T} = \frac{\Delta t}{T} = \frac{\beta}{\beta - 1} l_n \beta \quad \text{for } RC \neq T$$

and

$$t_m = \Delta t = T \quad \text{for } RC = T. \tag{10}$$

A curve of (t_m/T) as a function of β is shown in Fig. 12. From the relations of (9) and (10) the ratio of the

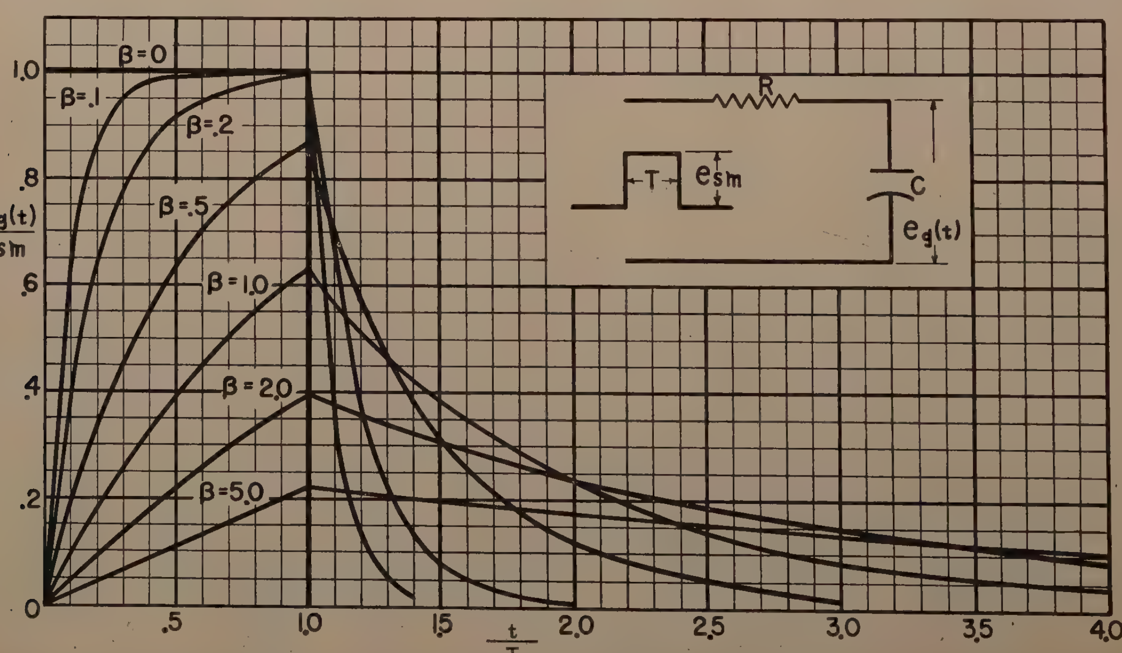

Fig. 9—The ratio $(e_g(t)/e_{sm})$ as a function of (t/T) for a square pulse with $\beta = RC/T$ as a parameter.

maximum grid voltage to the peak of the exponential pulse is

$$e_{gm}/e_{sm} = \frac{1}{1-\beta}\left[\beta^{(\beta/1-\beta)} - \beta^{(1/1-\beta)}\right] \text{ for } RC \neq T$$

$$= 1/\epsilon = 0.367 \qquad \text{for } RC = T. \qquad (11)$$

This ratio (e_{gm}/e_{sm}) is shown as a function of β in Fig. 12.

draws a current ranging from a few hundredths of a microampere up to several microamperes, or even higher for some of the larger, high-power thyratrons. Unless the grid resistance R is of the order of a megohm or more, the influence of this preconduction grid current for small, low-power thyratrons is small. However, if a large grid resistance is used, or if the tube has an exces-

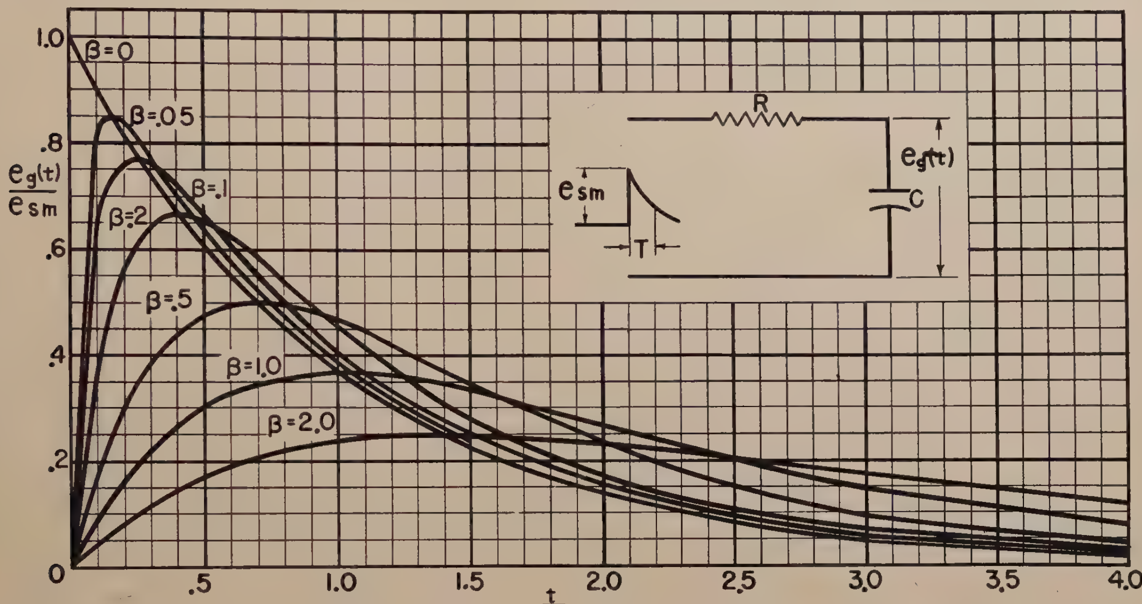

Fig. 11—The ratio $(e_g(t)/e_{sm})$ as a function of (t/T) for an exponential pulse with $\beta = RC/T$ as a parameter.

III. Application of the Analysis

General Discussion

The purpose of this paper is to present information from which the influence of the grid circuit on the grid-potential wave form can be predicted for several standard-signal wave forms. The validity of the foregoing analysis is based on the assumption that the thyratron presents a relatively high impedance to the grid circuit. Although the grid-cathode capacitance of the tube itself offers no particular difficulty, as it can be combined with the capacitance added externally to form the equivalent capacitance C used in the analysis,

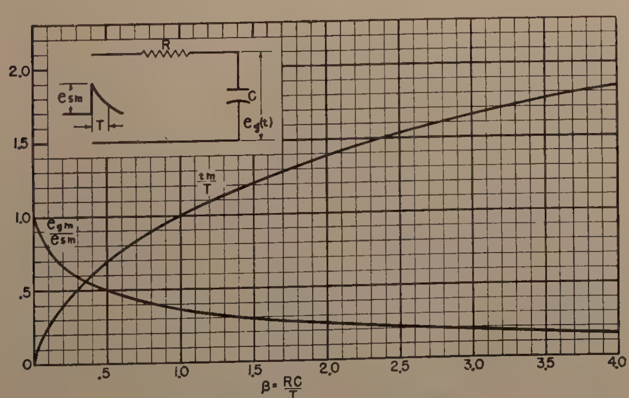

Fig. 12—The ratios (t_m/T) and (e_{gm}/e_{sm}) as functions of $\beta = RC/T$ for an exponential pulse.

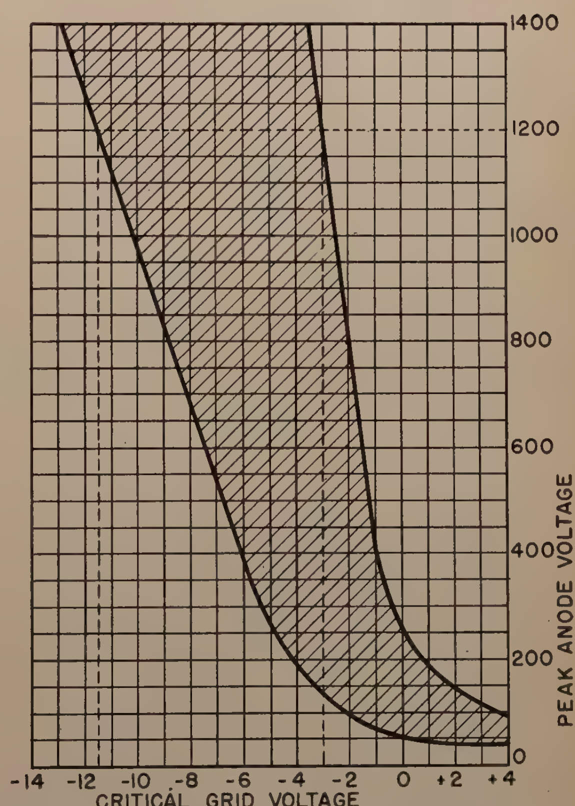

Fig. 13—Control band showing a possible range of variation of control characteristic for a thyratron.

the volt-ampere characteristic of the grid electrode does require consideration. It has been found that during the period prior to the initiation of the discharge, the grid

sive preconduction grid current, an elaboration of the analysis must be made to correct for the IR drop in the grid resistance. In many cases this grid current is

reasonably constant over a considerable range of negative grid voltage and the correction amounts to a shift in the direct-current bias value.

It is usually assumed that the thyratron fires at the instant the grid potential intersects the static grid-control characteristic. For simplicity we shall do likewise. This assumption, while quite satisfactory in most cases, is not strictly correct because of the time required to ionize the tube. Measurements indicate that this ionization time can range from a few microseconds down to very small fractions of a microsecond, depending upon the amount of grid over-voltage, the anode voltage, gas density, and circuit parameters. This complex dependence of the ionization time upon so many factors discourages its introduction into an analysis. Although this paper is concerned primarily with the influence of the grid-circuit parameters on the grid wave form during the preconduction period, a word should be said about the influence of tube conduction on the grid circuit. After initiation of the discharge, the grid input impedance decreases to an extremely small fraction of its preconduction value. During the period of tube conduction, the grid assumes a potential of a few volts positive with respect to the cathode (if the grid-circuit resistance is more than a few hundred ohms). This makes it possible to approximate the effect of tube conduction on the grid circuit by replacing the grid input impedance by a small positive grid-cathode potential of about 5 volts.

Relation of the Pulse Response to the Bias Voltage

The grid response has been given for only the signal pulse, whereas the presence of a grid bias voltage has not been considered. If the circuit elements are linear, the principle of superposition may be employed to combine the responses to the signal pulse and to the bias voltage. Thus, in the case of a direct-current bias, the pulse response may be added directly to the direct-current bias value. For an alternating-current bias, the response to this bias may be evaluated and added to the pulse response. In many cases the alternating-current bias voltage remains sensibly constant during the pulse width T, hence the pulse response may be added to the grid potential existing just prior to the application of the pulse.

Illustrative Example

A typical problem may best illustrate the use of the response curves for various pulse shapes and grid-circuit parameters. The following example indicates how these response curves may be used to estimate the degree of precision of firing for a specific application.

Most manufacturers of thyratrons present their grid-control information in the form of a control "band," as in Fig. 13. This band indicates the variation in control characteristics to be expected from such things as (1) variation in condensed-mercury temperature; (2) manufacturing variations from tube to tube; and (3) variations throughout the life of the tube.

Suppose a triangular signal pulse is applied to the circuit of Fig. 3 at an instant for which the anode voltage is 1200 volts. Reference to Fig. 13 shows that for an anode potential of 1200 volts, the critical grid potential may range from -3.0 to -11.5 volts. The application of the triangular pulse will cause the grid potential to sweep through this critical range in a time τ. This time interval τ may be regarded as a measure of the firing precision. Suppose the parameters of the grid circuit and the triangular pulse are as follows:

$$R = 50{,}000 \text{ ohms}$$
$$C = 0.004 \text{ microfarad}$$
$$e_{sm} = 150 \text{ volts}$$
$$T = 200 \text{ microseconds}$$
$$E_c = 50 \text{ volts.}$$

From this information $\beta=(RC/T)=1$. The response

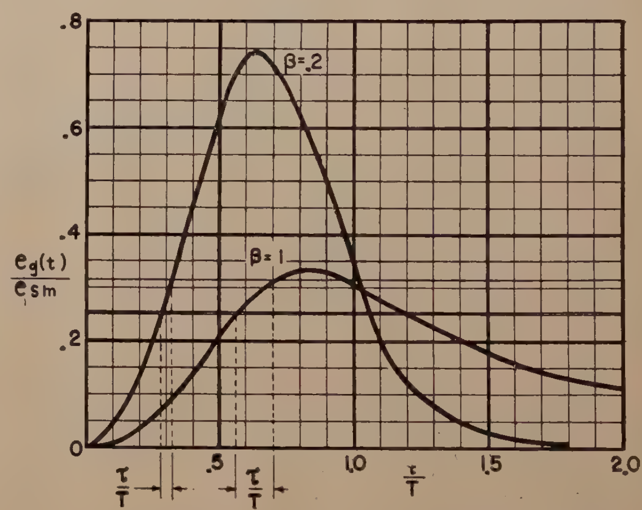

Fig. 14—Graphical determination of the precision of firing.

curve for a triangular pulse with $\beta=1$ is selected from Fig. 6 and repeated in Fig. 14. The critical grid voltages and the bias voltage are introduced in the following manner: assuming that the change in anode voltage is inappreciable during the time of the pulse duration T, the critical grid potentials are also constant over the interval of interest. Now, in order for the tube to fire, the grid potential must rise from a potential ($-E_c=-50$ volts) to the critical potential (ranging from -11.5 to -3.0 volts). Hence the pulse must raise the grid voltage by 47 volts to assure firing. To evaluate τ with the aid of the response curve it is necessary to "normalize" these voltage increments (38.5 and 47 volts) by dividing by the peak of the triangular pulse e_{sm}. These values of (e_g/e_{sm}), 0.313 and 0.258, are then laid out on the curve for $\beta=1$ in Fig. 14. From the resulting intersections the value of τ/T is found to be 0.13. For a 200-microsecond pulse, the value of τ is 26 microseconds. It should be noted that one intersection is only slightly below the peak of the response curve. Such a condition should be avoided as it would not insure dependable firing. The situation can be improved by decreasing the grid resistance to 10,000 ohms. The response curve for $\beta=0.2$ is then taken from Fig. 6 and laid out in Fig. 14. The intersections with this new curve result in a much better precision of firing, $\tau=9$ microseconds. In addition, the

response curve for $\beta=0.2$ rises considerably above $(e_g(t)/e_{sm})=0.313$ thus assuring an ample safety factor for dependable firing.

Mathematical Appendix

I. Response to a Triangular Pulse

The integrodifferential equation for a resistance-capacitance circuit is

$$Ri(t) + \frac{1}{C}\int_0^t i(t)dt + e_g(0) = e_s(t) \quad (12)$$

where $e_g(0)$ is the grid-capacitor voltage at $t=0$. For $e_g(0)=0$, the corresponding Laplace transform equation is

$$RI(p) + \frac{1}{Cp} T(p) = E_s(p), \quad (13)$$

from which

$$I(p) = \frac{1}{R}\frac{p}{p+1/RC} E_s(p) \quad (14)$$

and, since

$$E_g(p) = \frac{1}{CP} I(p) \quad (15)$$

the L transform of the capacitor voltage is

$$E_g(p) = \frac{1}{RC}\frac{1}{p+1/RC} E_s(p). \quad (16)$$

The L transform of the triangular signal pulse of Fig. 4 (a) is

$$E_s(p) = \frac{2e_{sm}}{p^2T}\left[1 - 2\epsilon^{-p(T/2)} + \epsilon^{-pT}\right]. \quad (17)$$

The three terms within brackets will be recognized as corresponding to the three time functions into which the triangular pulse may be divided. The two exponential terms are "shift" operators which shift the time response corresponding to $(2e_{sm}/p^2T)$ by $T/2$ and T, respectively. Substitution of (17) in (16) results in

$$E_g(p) = \frac{2e_{sm}}{RCT}\frac{1}{p^2(p+1/RC)}(1 - 2\epsilon^{-p(T/2)} + \epsilon^{-pT}). \quad (18)$$

The grid voltage as a time function, $e_g(t)$, is found by evaluating the inverse L transform of $E_g(p)$. This leads to the three expressions of (1) for $e_g(t)$.

II. Response to a Square Pulse

With no initial charge on the capacitor, the L transform of the grid voltage is again

$$E_g(p) = \frac{1}{RC}\frac{1}{p+1/RC} E_s(p). \quad (16)$$

The L transform for a square pulse of height e_{sm} and duration T is

$$E_s(p) = \frac{e_{sm}}{p}(1 - \epsilon^{-pT}), \quad (19)$$

where the two te ms in brackets correspond to the two step-functions which combin to form the square pulse. Upon substitution of (19) in (16), the L transform of the grid voltage becomes

$$E_g(p) = \frac{e_{sm}}{RC}\frac{1}{p(p+1/RC)}(1 - \epsilon^{-pT}). \quad (20)$$

As before, the exponential term is a "shift" operator which shifts the response to a unit step-function by the time T. The inverse L transform, $e_g(t)$, is given by (6).

III. Response to an Exponential Pulse

As in the preceding cases, the L transform of the grid voltage is

$$E_g(p) = \frac{1}{RC}\frac{1}{(p+1/RC)} E_s(p), \quad (16)$$

where the L transform of an exponential pulse of initial height e_{sm} and time constant T is

$$E_s(p) = \frac{e_{sm}}{p+1/T}. \quad (21)$$

With the substitution of (21) in (16), the L transform of the grid voltage is

$$E_g(p) = \frac{e_{sm}}{RC}\frac{1}{(p+1/RC)(p+1/T)}. \quad (22)$$

This L transform leads to two time functions: one for the general case $RC\neq T$, and one for the special case $RC=T$. The corresponding time functions for $e_g(t)$ are given by (9).

Acknowledgment

We wish to thank Mr. D. E. Marshall, of the Industrial Tube Section of this company, for his encouragement and helpful comments during the preparation of this paper.

The Steady-State Operational Calculus*

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Summary—**The direct and inverse transforms of the steady-state operational calculus are presented, together with two methods of evaluating the inverse transform, the first resulting in a Fourier series and the second giving a sum function. A proof of the inversion theorem connecting the two transforms is outlined in the Appendix. Two examples are presented illustrating the application of this operational calculus to circuit problems, and a comparison is made between the ordinary and the steady-state operational calculuses.**

Introduction

THE USE of nonsinusoidal wave forms in electric circuits has increased greatly in the past few years, particularly with the advent of new and improved electronic devices. Fourier series may be used in the solution of the circuit problems that arise in these applications if the harmonic voltages and currents with appreciable amplitude are few in number. If, however, the harmonics with large amplitudes are great in number, other methods of circuit solution must be utilized. These methods of steady-state solution of circuits with nonsinusoidal wave forms recently have been the object of increased study, as evidenced by the number of investigations that have been made.[1] It appears that many of the results obtained in these scattered investigations may be integrated into a steady-state operational calculus which will present a more consistent method of solving for the steady state of nonsinusoidal circuit problems. It is the purpose of this paper to present such a steady-state operational calculus; an operational calculus in many ways quite similar to and yet in other ways quite different from the ordinary operational calculus. The method will also be applied to several examples.

An operational calculus is a method of solving differential equations in which the process of integration and differentiation is replaced by an operator which is manipulated as an algebraic quantity. The resulting operational equation is solved, and the operational solution is then transformed into the actual solution of the differential equation. As a result of this method, the ordinary operational calculus produces a solution of a differential equation of, for example, a current, which is composed of a transient part and a steady-state part. These two parts are difficult to identify and separate in many cases, and the steady-state part is seomtimes in the form of a Fourier series which is hard to use and interpret. As a result of the steady-state operational calculus, the steady-state current is obtained by itself, and the current may be expressed either as a Fourier series or as a sum function. In a given circuit with a nonsinusoidal periodic voltage applied, the current may be obtained by changing the voltage into a Fourier series, calculating the current for the fundamental and each harmonic, and adding the results to obtain the Fourier series of the current. To obtain the wave form of the current from this series is very tedious, particularly if the voltage has any sharp discontinuities in it. The current, on the other hand, may be expressed as a sum function of the series by the use of the steady-state operational calculus, and the current wave form is easily plotted from this sum function.

The operational method, in brief, consists of operating on the impressed voltage by means of the direct transform. The transformed voltage is divided by the operational impedance to obtain the transformed current. The transformed current is then operated upon by the inverse transform to obtain the resulting current. The two most important concepts of any operational calculus are the direct and the inverse transforms, and these will be introduced first.

Transforms

The direct transform was introduced by McLachlan[2] in 1937. If the periodic time function $f(t)$ has a period of T seconds, then the direct transform is

$$S[f(t)] = \int_0^T \epsilon^{-pt} f(t) dt \tag{1}$$

where p is a complex number. The inverse transform has been introduced[3] recently, and was used to obtain the steady-state solution of circuits. It is

$$S^{-1}[F(p)] = \frac{1}{2\pi j} \int_W \frac{\epsilon^{pt} F(p)}{(1 - \epsilon^{-pT})} dp \tag{2}$$

where $F(p)$ is a function of p with properties as outlined in the Appendix. It is possible to extend this inverse transform to include the case of functions with branch points such as those encountered in transmission-line theory, but discussion of this more difficult integral will be deferred for the present. The type of function considered here will contain practically all of the cases encountered in ordinary linear-circuit theory using concentrated parameters. For the inverse transform, the path of integration W in the complex plane is composed of two parts, W_1 and W_2 as shown in Fig. 1. All of the poles of $F(p)$ must be to the left of W_2, except for the points, $p = (jn2\pi/T) = jn\omega$ where n is a positive or negative integer, which must lie between W_1 and W_2. The

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[1] See List of Additional References and footnotes 2, 3, and 4.

[2] N. W. McLachlan, "Fourier expansions obtained operationally," *Phil. Mag.*, vol. 24, pp. 1055–1058; December, 1937.

[3] D. L. Waidelich, "Steady-state currents of electrical networks," *Jour. Appl. Phys.*, vol. 13, pp. 706–712; November, 1942.

path W_3 to be used later is the same as W_2 except that the direction is reversed.

It is explained in the outline of the inversion-theorem proof as given in the Appendix that the inverse transform is unique; i.e., it has one and only one value for a given value of t, whereas the direct transform is not unique. This is one of the points of difference between the ordinary operational calculus and the operational method outlined in this paper. There is one value of the direct transform that is unique, however, and that is the transform $F(p)$ which has no poles anywhere in the finite part of the complex plane. It is also the transform obtained always from (1) and will henceforth be called the primitive direct transform.

The evaluation of the direct transform may be accomplished readily by treating the operator p as a constant, and by using an ordinary table of integrals. To evaluate the inverse transform, on the other hand, requires a working knowledge of the theory of residues for functions of a complex variable.[4] The evaluation of the inverse transform has been treated in detail,[3] so only a summary of the method will be given here.

1. *Evaluation as a Fourier series:* The residues of the integral at the poles $p=(jn2\pi/T)$ within the path of integration W are calculated and summed to give the Fourier series.

2. *Evaluation as a sum function:* In many cases, the inverse transform may be evaluated as the sum function of the Fourier series. The transform may be expressed as two separate integrals

$$S^{-1}[F(p)] = \frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}dp$$
$$= \frac{1}{2\pi j}\int_{W_1} \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}dp - \frac{1}{2\pi j}\int_{W_3} \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}dp. \quad (3)$$

The path of integration W_1 of the first integral of (3) is the same as that of the inverse transform of the ordinary operational calculus and may be evaluated in the same manner.[3] Thus

$$\frac{1}{2\pi j}\int_{W_1} \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}dp$$
$$= \frac{1}{2\pi j}\int_{W_1} \epsilon^{pt}F(p)dp + \frac{1}{2\pi j}\int_{W_1} \epsilon^{p(t-T)}F(p)dp + \cdots, \quad (4)$$

and when $0<t<T$,

$$\frac{1}{2\pi j}\int_{W_1} \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}dp = \frac{1}{2\pi j}\int_{W_1} \epsilon^{pt}F(p)dp \quad (5)$$

which may be evaluated by calculating the residues at all poles to the left of W_1. The second integral of (3) may be evaluated by calculating the residues at the poles to the left of W_3 since the contribution of the integral around a large semicircle to the left of W_3 can be shown

Fig. 1—One path of integration employed in the inverse transform.

to approach zero as the radius of the semicircle approaches infinity. Difficulty is sometimes encountered when some of the poles of $F(p)$ lie on the imaginary axis or to the right of the imaginary axis. In these cases, the paths of integration W_2 or W_3 must be bent around these poles so that the poles are to the left of the paths. This will be illustrated in the examples of the next paragraph.

The first example will be that of a sine wave $e(t)=E_m \sin(\omega t+\phi)$. From (1) the direct transform is

$$S[e(t)] = \int_0^T \epsilon^{-pt}E_m \sin(\omega t+\phi)dt$$
$$= \frac{E_m(p\sin\phi + \omega\cos\phi)(1-\epsilon^{-pT})}{p^2+\omega^2}. \quad (6)$$

The inverse transform is obtained by substituting (6) into (2)

$$S^{-1} = \frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}E_m(p\sin\phi+\omega\cos\phi)}{p^2+\omega^2}dp$$
$$= \frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}E_m(p\sin\phi+\omega\cos\phi)}{(p-j\omega)(p+j\omega)}dp. \quad (7)$$

The integral (7) has two poles $p=\pm j\omega$, both of which lie within the path of integration W and hence the residue must be found at both poles. When this is done it will be found that

$$S^{-1} = E_m \sin(\omega t+\phi). \quad (8)$$

Fig. 2 is an illustration of the full-wave rectified sine wave to be used as the second example.

$$S = \int_0^T \epsilon^{-pt}E_m \sin(\pi t/T)dt = \frac{E_m(\pi/T)(1+\epsilon^{-pT})}{p^2+(\pi^2/T^2)}. \quad (9)$$

The inverse transform

[4] N. W. McLachlan, "Complex Variable and Operational Calculus," Cambridge University Press, New York, N. Y., 1939.

$$S^{-1} = \frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}E_m(\pi/T)(1+\epsilon^{-pT})}{[p^2+(\pi^2/T^2)](1-\epsilon^{-pT})}dp \quad (10)$$

is evaluated as a Fourier series by calculating the residues at the poles $p=(jn2\pi/T)$ where n is any positive or negative integer.

$$S^{-1} = \frac{2E_m}{\pi}\sum_{n=-\infty}^{+\infty}\frac{\epsilon^{(j2\pi nt/T)}}{1-4n^2}$$

$$= \frac{2E_m}{\pi} - \frac{4E_m}{\pi}\sum_{n=1}^{\infty}\frac{\cos(2\pi nt/T)}{4n^2-1}. \quad (11)$$

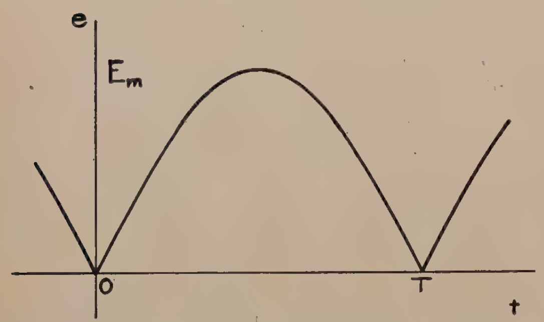

Fig. 2—The full-wave rectified sine-wave voltage.

Equation (10) may also be evaluated as the sum function of the Fourier series

$$S^{-1} = \frac{1}{2\pi j}\int_{W_1}\frac{\epsilon^{pt}E_m(\pi/T)(1+\epsilon^{-pT})(1+\epsilon^{-pT}+\epsilon^{-p2T}+\cdots)}{p^2+(\pi^2/T^2)}dp$$
$$-\frac{1}{2\pi j}\int_{W_3}\frac{\epsilon^{pt}E_m(\pi/T)(1+\epsilon^{-pT})}{[p^2+(\pi^2/T^2)](1-\epsilon^{-pT})}dp.$$

When $0<t<T$

$$S^{-1} = \frac{1}{2\pi j}\int_{W_1}\frac{\epsilon^{pt}E_m(\pi/T)}{p^2+(\pi^2/T^2)}dp$$
$$-\frac{1}{2\pi j}\int_{W_3}\frac{\epsilon^{pt}E_m(\pi/T)(1+\epsilon^{-pT})}{[p^2+(\pi^2/T^2)](1-\epsilon^{-pT})}dp. \quad (12)$$

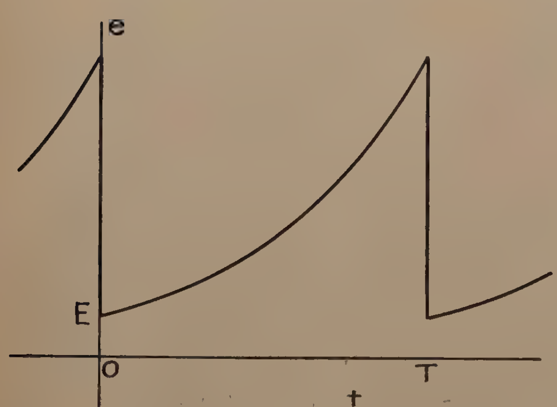

Fig. 3—Exponential wave form.

The first integral of (12) has two poles $p=\pm j(\pi/T)$, while the second has no poles; hence

$$S^{-1} = E_m \sin(\pi t/T). \quad (13)$$

The third example is that of the exponential wave form shown in Fig. 3 whose equation for $0<t<T$ is $e=E\epsilon^{at}$ where $a>0$.

$$S = \int_0^T \epsilon^{-pt}E\epsilon^{at}dt = \frac{E}{p-a}[1-\epsilon^{-(p-a)T}]. \quad (14)$$

The inverse transform as the sum function is

$$S^{-1} = \frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}E[1-\epsilon^{-(p-a)T}]}{(p-a)(1-\epsilon^{-pT})}dp$$
$$= \frac{1}{2\pi j}\int_{W_1}\frac{\epsilon^{pt}E}{p-a}dp$$
$$-\frac{1}{2\pi j}\int_{W_3}\frac{\epsilon^{pt}E[1-\epsilon^{-(p-a)T}]}{(p-a)(1-\epsilon^{-pT})}dp$$
$$= E\epsilon^{at}. \quad (15)$$

The first integral of (15) has one pole $p=a$, but the second integral has no poles. Determine now for the direct transform

$$S = \frac{E(1-\epsilon^{aT})}{p-a} \quad (16)$$

the corresponding inverse transform

$$S^{-1} = \frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}E(1-\epsilon^{aT})}{(p-a)(1-\epsilon^{-pT})}dp$$
$$= \frac{1}{2\pi j}\int_{W_1}\frac{\epsilon^{pt}E(1-\epsilon^{aT})}{p-a}dp$$
$$-\frac{1}{2\pi j}\int_{W_3}\frac{\epsilon^{pt}E(1-\epsilon^{aT})}{(p-a)(1-\epsilon^{-pT})}dp. \quad (17)$$

The path W_3 has to be bent as shown in Fig. 4 because the pole $p=a$ must be to the left of W_3. Both integrals

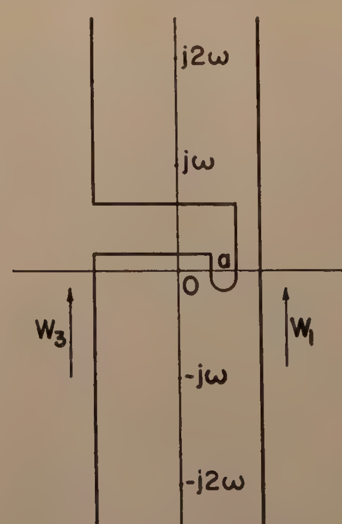

Fig. 4—An alternate path of integration employed in the inverse transform.

of (17) have the pole $p=a$, and when evaluated at this

pole, the inverse transform becomes

$$S^{-1} = E\epsilon^{at}. \tag{18}$$

It will be noticed that two different direct transforms, those of (14) and (16), both have the same inverse transform, indicating again that the direct transform is not unique. The primitive direct transform for this example is that of (14), since first, it has no singularities in the finite part of the complex plane, and second, it is obtained from (1). Equation (16) is one of the infinite number of nonprimitive transforms for this example, and was obtained from the primitive transform by subtracting the quantity

$$\frac{E\epsilon^{aT}(1 - \epsilon^{-pT})}{p - a} \tag{19}$$

which, when substituted in the inverse transform, will be found to be zero. Just as there are two forms of the inverse transform; i.e., the Fourier series and the sum function, there are also two forms of the direct transform, the series form and the sum-function form. From the second part of the inversion theorem of the Appendix, if $F(p)$ is a primitive direct transform, then

$$F(p) = \sum_{n=-\infty}^{+\infty} \frac{(1 - \epsilon^{-pT})F(j2\pi n/T)}{T(p - j2\pi n/T)} \cdot \tag{20}$$

As an illustration, the primitive transform (14) of example 3 may be expanded into the series

$$S = \frac{E}{p - a}\left[1 - \epsilon^{-(p-a)T}\right]$$
$$= \sum_{n=-\infty}^{+\infty} \frac{(1 - \epsilon^{-pT})E(\epsilon^{aT} - 1)}{T(p - j2\pi n/T)(a - j2\pi n/T)} \cdot \tag{21}$$

Circuit Analysis

Three simple theorems are necessary before this method may be applied to the solution of steady-state circuit problems. These are

1. If $f_n(t) = S^{-1}[F_n(p)]$, $f(t) = S^{-1}[F(p)]$, and $F(p) = \sum_{n=1}^{n} F_n(p)$, then $f(t) = \sum_{n=1}^{n} f_n(t)$.

This theorem may be inverted if $F(p)$ and $F_n(p)$ are primitive transforms.

2. If $g(t) = g(0) + \int_0^t f(t)dt$ and $g(t)$ is continuous, then

$$S[g(t)] = (1/p)(1 - \epsilon^{-pT})g(0) + (1/p)S[f(t)]. \tag{22}$$

The proof of this particular theorem will be carried through.

$$S[g(t)] = \int_0^T \epsilon^{-pT}g(0)dt + \int_0^T \epsilon^{-pt}\left[\int_0^t f(t)dt\right]dt$$
$$= (1/p)(1 - \epsilon^{-pT})g(0) + (1/p)\int_0^T \epsilon^{-pt}f(t)dt$$
$$= (1/p)(1 - \epsilon^{-pT})g(0) + (1/p)S[f(t)].$$

3. If $g(t) = d[f(t)]/dt$ and $f(t)$ is continuous, then

$$S[g(t)] = -(1 - \epsilon^{-pT})f(0) + pS[f(t)]. \tag{23}$$

Both the second and third theorems may be applied as many times as is necessary. A number of other interesting but less essential theorems may be derived in a similar manner, but these are not included here.

Two examples of the application of this steady-state operational calculus to the solution of circuit problems will be given.

1. The first example is that of the sinusoidal voltage $e(t) = E_m \sin(\omega t + \phi)$ applied to a series resistance R, inductance L, and capacitance C circuit, whose voltage equation is

$$e(t) = Ri + L\frac{di}{dt} + \frac{q}{c} \cdot \tag{24}$$

Let $I = S[i(t)]$, and by the use of the three preceding theorems, (24) becomes

$$S[e(t)] = RI - (1 - \epsilon^{-pT})Li(0) + LpI$$
$$+ (1/p)(1 - \epsilon^{-pT})[q(0)/c] + (I/pC). \tag{25}$$

Solving for I,

$$I = \frac{pS[e(t)]}{L(p^2 + pR/L + 1/LC)} + \frac{p(1 - \epsilon^{-pT})i(0)}{p^2 + pR/L + 1/LC}$$
$$- \frac{(1 - \epsilon^{-pT})q(0)}{LC(p^2 + pR/L + 1/LC)} \cdot \tag{26}$$

The expression for I in (26) is the primitive transform of the steady-state current $i(t)$. The last two terms of (26), when substituted into the inverse transform, will be zero, and hence the first term is the only one that need be used. This first term is a nonprimitive transform of $i(t)$, and in effect its use means that (22) and (23) become

$$S\left[g(0) + \int_0^t f(t)dt\right] = (1/p)S[f(t)] \tag{27}$$

and

$$S[df(t)/dt] = pS[f(t)]. \tag{28}$$

These two equations can be used for almost all circuit problems, but if there is any doubt, (22) and (23) should be used. Using the first term of (26) and with the aid of (2) and (6),

$$i = \frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}E_m p(p \sin\phi + \omega\cos\phi)}{L(p^2 + \omega^2)(p^2 + pR/L + 1/LC)}dp$$
$$= (E_m/Z)\sin(\omega t + \phi - \theta) \tag{29}$$

where

$$Z = \sqrt{R^2 + (\omega L - 1/\omega C)^2} \quad \text{and} \quad \theta = \tan^{-1}[(\omega L - 1/\omega C)/R].$$

This is the familiar expression for the steady-state current of this circuit.

2. The square-wave voltage used in communications work is shown in Fig. 5 and has the direct transform

$$S[e(t)] = (E_m p)(1 - \epsilon^{-pT/2})^2. \tag{30}$$

The voltage is applied to the circuit of Fig. 6, and the voltage e_c across the capacitance C is required. The

transform of this voltage is

$$E_C = \frac{pLS[e(t)]}{p^2CRL + pL + R}, \tag{31}$$

and the inverse transform of (31), for $0<t<T/2$, is

$$e_c = 4E_m\left(\frac{m}{n}\right)\epsilon^{-nt}\left[\frac{\sin nt + \epsilon^{-mT/2}\sin n(t-T/2)}{1+2\epsilon^{-mT/2}\cos(nT/2)+\epsilon^{-mT}}\right] \tag{32}$$

where $m=(1/2CR)$, $n=\sqrt{(1/LC)-m^2}$, and $LC<(2CR)^2$. The voltage e_c is shown in Fig. 7 for the values

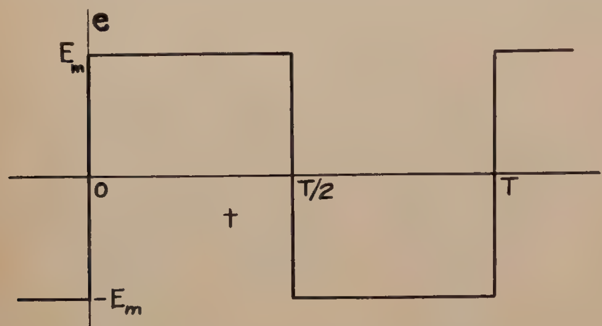

Fig. 5—The square-wave voltage applied to the circuit of Fig. 6.

$\omega^2LC=0.236$ and $\omega CR=1.0$ where $\omega=(2\pi/T)$. When $LC>(2CR)^2$,

$$e_c = 2E_m\left(\frac{a+b}{a-b}\right)\left[\frac{\epsilon^{-bt}}{1+\epsilon^{-bT/2}} - \frac{\epsilon^{-at}}{1+\epsilon^{-aT/2}}\right] \tag{33}$$

where $a=m+jn$ and $b=m-jn$. The wave form of Fig. 6 may also be observed experimentally.

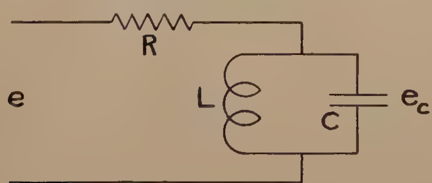

Fig. 6—Resistance-inductance-capacitance circuit.

A table of corresponding direct and inverse transforms would be very useful in solving for the steady-state currents of circuits, but such a table would have to be very lengthy because of the many different types of voltages that are encountered in this work, and also because of the many direct transforms, both the primitive one and nonprimitive ones, that correspond to a given inverse transform. For this reason, a compilation of a transform table was not attempted here, but it is hoped to present one later.

A succinct comparison of the ordinary operational calculus and the steady-state operational calculus is presented in Table I. It will be noticed that in both operational calculuses, the derivative sign is replaced by p and the integral sign by $(1/p)$.

Conclusions

The direct and inverse transforms of the steady-state operational calculus have been presented, together with an inversion theorem that shows that the two transforms are actually inverses of each other. With the aid of several theorems involving the derivative of a function and the integral of a function, it was demonstrated that the inverse transform, when evaluated, yields the steady-state currents and voltages of the circuit under study. This method was then applied to two examples,

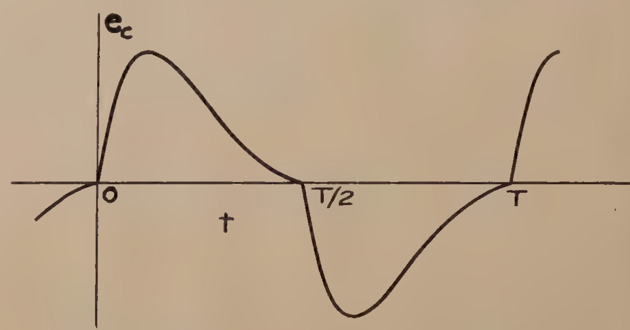

Fig. 7—Resulting voltage e_c across the capacitance.

and the solutions were obtained as sum functions which are very convenient in plotting the wave forms involved. On the other hand, the solutions may be obtained as a Fourier series if the values of the harmonics are desired.

Table I

	Ordinary Operational Calculus	Steady-State Operational Calculus
Direct Transform	$L[f(t)] = \int_0^\infty \epsilon^{-pt} f(t)\,dt$	$S[f(t)] = \int_0^T \epsilon^{-pt} f(t)\,dt$
Inverse Transform	$L^{-1}[F(p)] = \frac{1}{2\pi j}\int_{c-j\infty}^{c+j\infty} \epsilon^{pt}F(p)\,dp$ where c is greater than the real parts of the poles of $F(p)$.	$S^{-1}[F(p)] = \frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}\,dp$
Transform of a Derivative	$L\left[\frac{df(t)}{dt}\right] = pL[f(t)]$ if $f(0)=0$	$S\left[\frac{df(t)}{dt}\right] = pS[f(t)]$
Transform of an Integral	$L\left[\int_0^t f(t)\,dt\right] = (1/p)L[f(t)]$	$S[g(t)] = (1/p)S[f(t)]$ where $g(t) = g(0) + \int_0^t f(t)\,dt$

Appendix

Outline of Inversion-Theorem Proof

Part 1: It will be assumed here, as is true in almost all practical problems, that $f(t)$ is bounded and has at most a finite number of discontinuities.

If $F(p)=\int_0^T \epsilon^{-pt}f(t)dt$, then

$$f(t) = \frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}\,dp.$$

For if $\alpha>0$ and $y>0$, let

$$A_1 = \frac{1}{2\pi j}\int_{\alpha-jy}^{\alpha+jy} \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}\,dp$$

$$= \frac{1}{2\pi j}\int_{\alpha-jy}^{\alpha+jy} dp \int_0^T \frac{\epsilon^{p(t-x)}f(x)}{1-\epsilon^{-pT}}\,dx$$

$$= \frac{1}{2\pi j}\int_0^T dx \int_{\alpha-jy}^{\alpha+jy} \frac{\epsilon^{p(t-x)}f(x)}{1-\epsilon^{-pT}}\,dp$$

$$= \frac{1}{2\pi j}\int_0^T dx \int_{\alpha-jy}^{\alpha+jy} f(x)\epsilon^{p(t-x)}\left[\sum_{n=0}^{\infty}\epsilon^{-npT}\right]dp$$

$$= \frac{1}{2\pi j}\int_0^T f(x)\left\{\sum_{n=0}^{\infty}\int_{\alpha-jy}^{\alpha+jy}\epsilon^{p(t-x-nT)}dp\right\}dx$$

$$= \frac{1}{\pi}\int_0^T f(x)\left\{\sum_{n=0}^{\infty}\frac{\epsilon^{\alpha(t-x-nT)}\sin y(t-x-nT)}{t-x-nT}\right\}dx$$

$$= \sum_{n=0}^{\infty}\frac{1}{\pi}\int_{t-(n+1)T}^{t-nT} f(t-nT-u)\left(\frac{\epsilon^{\alpha u}\sin yu}{u}\right)du.$$

Similarly, if $\beta>0$ and $y>0$, let

$$A_2=\frac{1}{2\pi j}\int_{-\beta-jy}^{-\beta+jy}\frac{\epsilon^{pt}F(p)}{1-\epsilon^{-pT}}dp$$

$$=-\sum_{n=0}^{\infty}\frac{1}{\pi}\int_{t+nT}^{t+(n+1)T} f[t+(n+1)T-u]\frac{\epsilon^{-\beta u}\sin yu}{u}du.$$

Then

$$\frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}dp$$

$$= \lim_{y\to\infty}(A_1-A_2)$$

$$=(1/2)\left[f(t+0)+f(t-0)\right] \text{ if there is a discontinuity at } t.$$

Or

$$\frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}dp = f(t) \text{ if there is no discontinuity.}$$

It is also possible to show that the inverse transform is unique.

Part 2. It will be assumed here that $F(p)$ is a meromorphic function of p; i.e., a function analytic everywhere in the finite part of the complex plane except for a finite number of poles as singularities in each finite part of the complex plane. The function $F(p)$ must not have any essential singularities or branch points in any finite portion of the plane.

If

$$f(t) = \frac{1}{2\pi j}\int_W \frac{\epsilon^{pt}F(p)}{(1-\epsilon^{-pT})}dp$$

and

$$\left|\frac{F(z)}{1-\epsilon^{-zT}}\right| \to 0 \text{ as } r\to\infty \text{ where } z=\frac{(2r+1)\pi}{T}\epsilon^{j\theta}$$

and r is an integer,

then $F(p)=\int_0^T\epsilon^{-pt}f(t)dt$.

For if $\alpha>0$ and $y>0$, let

$$B_1=\int_0^T \epsilon^{-pt}\left\{\frac{1}{2\pi j}\int_{\alpha-jy}^{\alpha+jy}\frac{\epsilon^{zt}F(z)}{(1-\epsilon^{-zT})}dz\right\}dt$$

where all of the singularities of $F(z)$ lie to the left of the line $z=\alpha$.

$$B_1=\frac{1}{2\pi j}\int_{\alpha-jy}^{\alpha+jy}\frac{F(z)}{(1-\epsilon^{-zT})}\left\{\int_0^T\epsilon^{(z-p)t}dt\right\}dz$$

$$=\frac{1}{2\pi j}\int_{\alpha-jy}^{\alpha+jy}\frac{F(z)\left[1-\epsilon^{-(p-z)T}\right]}{(p-z)(1-\epsilon^{-zT})}dz.$$

Similarly, if $\beta>0$ and $y>0$, let

$$B_2=\int_0^T \epsilon^{-pt}\left\{\frac{1}{2\pi j}\int_{-\beta-jy}^{-\beta+jy}\frac{\epsilon^{zt}F(z)}{(1-\epsilon^{zT})}dz\right\}dt$$

$$=\frac{1}{2\pi j}\int_{-\beta-jy}^{-\beta+jy}\frac{F(z)\left[1-\epsilon^{-(p-z)T}\right]}{(p-z)(1-\epsilon^{-zT})}dz$$

where all the singularities of $F(p)$ are to the left of the line $p=-\beta$ except for the poles $p=j2\pi m/T$. Consider the integral

$$D_m=\frac{1}{2\pi j}\int_{C_m}\frac{F(z)\left[1-\epsilon^{-(p-z)T}\right]}{(p-z)(1-\epsilon^{-zT})}dz$$

where m is a positive integer and C_m is a rectangular curve traversed counterclockwise and with straight-line sides, whose equations are $p=\alpha$, $p=j(2m+1)\pi/T$, $p=-\beta$, and $p=-j(2m+1)\pi/T$. Then

$$D_m=\sum_{m=-m}^{+m} G_m\left(\frac{1}{p-j\dfrac{2\pi m}{T}}\right)$$

where G_m is the principal part at the pole $p=j2\pi m/T$. The contribution to D_m of the integrals along the lines $p=\pm j(2m+1)\pi/T$ approaches zero as $m\to\infty$. Hence

$$\int_0^T\epsilon^{-pt}f(t)dt=\lim_{y\to\infty}(B_1-B_2)=\lim_{m\to\infty}D_m$$

$$=\sum_{m=-\infty}^{+\infty}G_m\left(\frac{1}{p-j2\pi m/T}\right)$$

$$=F(p)+H(p)$$

where $H(p)$ is a function of p which, when substituted in the inverse transform, will turn out to be zero. Hence the direct transform is not unique but may have any number of values.

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Antinoise Characteristics of Differential Microphones*

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Summary—**The operation of the differential[1,2] microphone is described, and the noise discrimination of differential microphones operating on various orders of pressure gradients is obtained. Equations of the discrimination against random sounds are developed for both theoretical and practical microphones. Curves are drawn, showing discrimination versus frequency. The noise discrimination of the nth-order pressure gradient is obtained as the ultimate in differential-type microphones.**

I. Introduction

AMONG the many refinements in the technique of electrical transmission of sound, a substantial number has been concerned with exclusion from the transmitter of unwanted, ambient noise arising from extraneous sources and of reflections of the desired sound. The problem has been met satisfactorily in cases where acoustic conditions at the transmitting location can be controlled by suitable insulation against undesired sounds and by treatment of the enclosing walls to reduce or diffuse reflections. These methods, however, give no relief to the difficulties encountered when a communication microphone must be located in a zone of high ambient noise. While baffles, shields, directional microphones, weighted response curves, throat microphones,[3] etc., are sometimes helpful, none provides sufficiently complete noise rejection to outweigh their various inherent disadvantages, such as loss of articulation and intelligibility, large, awkward, or uncomfortable microphones, or limitations on movement of the speaker. This problem became really serious when the communication needs of modern, mechanized warfare required direct-speech communication from such locations as the inside of a tank, where ambient noise commonly runs to levels beyond any peacetime precedent. The United States Army Type T-45 lip microphone, developed by Burroughs and Kahn from the original conception of Beekley,[2] has proven a thoroughly satisfactory solution of the military problem; and other versions of the same microphone, adapted to hand-held, handset, and switchboard-operator types, promise important improvement in many peacetime telephone and radiophone communication services.

Considering the differential[1] microphone as a variety of pressure-gradient microphone, the characteristics of it may be examined mathematically by methods similar to those that have been applied to other pressure-gradient microphones of heretofore familiar varieties.[4–7]

One of the characteristics of a differential microphone is the difference in frequency response and level when actuated by a sound wave from a close source, over that of a distant source. The response of a pressure-gradient microphone to a spherical wave is a function of the distance to the sound source. Most users of velocity microphones are familiar with this phenomenon, as the microphone accentuates the low frequencies when the microphone is brought nearer the sound source. Another characteristic of the pressure-gradient microphone is its polar response curve which is a function of the cosine of the angle which the incident sound wave makes with the microphone. This polar characteristic reduces the response of the microphone for random sounds. The difference in response and level due to the proximity of the sound source combined with its polar characteristics makes the pressure-gradient microphone ideally suited for an antinoise microphone.

It is the purpose of this paper to show how this proximity effect and polar response may be utilized to produce an antinoise and antifeedback microphone.

II. Theoretical First-Order Differential Microphone

The basic principle of the differential microphone may be clearly represented by plotting pressure versus distance in a spherical wave as the wave progresses from a point source. Since the pressure in a spherical wave varies inversely as the distance from the source, a differential microphone operates on a rapidly decreasing portion of the pressure curve if the microphone is close to the sound source, while if the microphone is a considerable distance from the sound source the change in pressure, in a distance equal to the spacing between the sound entrances, is small as shown in Fig. 1.

While Fig. 1 shows that the difference of sound pressures at the two sound entrances is large for short distances from the source and practically zero for large distances, this figure is altered by the fact that phase change of the pressure wave as a function of distance has been ignored. There will be a change in phase caused by any difference in acoustic distance from the sound

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[1] The following definition has been used for "differential microphone" in this paper: "A differential microphone is a pressure-gradient microphone which has two or more entrances which are separated by an acoustic distance which is small compared to the wavelength for the sound which may impinge on opposite sides of the same diaphragm or on separate diaphragms, so that the resultant force actuating the microphone for a distant sound source is attenuated, while the resultant force for a close sound source is preponderantly that resulting from the sound pressure at the openings nearer the sound source.

[2] F. C. Beekley, United States Patent No. 2,350,010.

[3] J. Shawn, "Application of the throat microphone," *Communications*, vol. 23, p. 11; January, 1943.

[4] H. F. Olson, "Mass controlled electrodynamic microphones: the ribbon microphone," *Jour. Acous. Soc. Amer.*, vol. 3, pp. 56–68; July, 1931.

[5] B. F. Miessner, United States Patent No. 1,507,081, describes a pressure-gradient microphone used as an antinoise microphone.

[6] B. Baumzweiger, United States Patent No. 2,237,298, describes a higher-order gradient microphone using two ribbons for increasing the directivity over that of a first-order gradient.

[7] H. F. Olson, United States Patent No. 2,301,744, describes several higher-order gradient microphones used for increased directivity over that of a first-order gradient microphone. The effect of a spherical wave on the response is also described.

source of the two sound entrances. Consequently, in developing equations for the differential microphone the change in magnitude and phase must be considered.

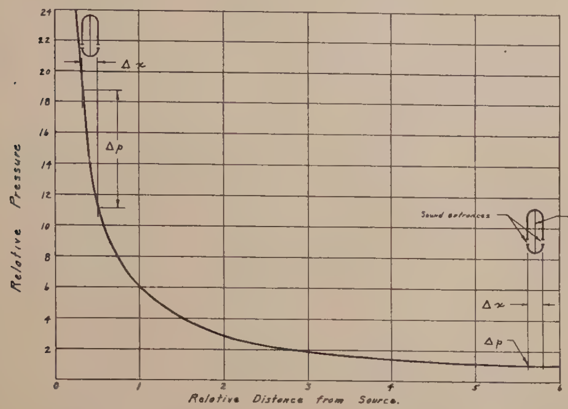


Fig. 1—Variation of pressure with distance from a point source.

In the discussion of differential microphones, there are two distinct cases that could be considered. The first is the comparison of the response of an ideal gradient microphone, which has infinitesimal thickness, to a sound source at a small distance from the microphone with the response to the same sound source when it is at a large distance from the microphone. The second and more general case, which includes the first case, is the response of the microphone to a source close to the microphone compared with the response to a sound source which is a large distance away but is producing the same pressure, at one surface of the microphone as the close sound source. It is the second case which will be analyzed, considering the sound sources as being on the axis of the microphone. Fig. 2 shows the physical setup for such a case.

P_1 = maximum pressure of sound source S_1 at unit distance from source.
P_2 = maximum pressure of sound source S_2 at unit distance from source.
r_1 = distance of sound source S_1 from microphone.
r_2 = distance of sound source S_2 from microphone.
p = instantaneous pressure at one side of microphone.
x = variable distance.
k = $2\pi/\lambda$.
λ = wavelength.
c = velocity of sound.
A = maximum pressure P_1 or P_2, as the case may be.
dx = infinitesimal distance between sound entrances of theoretical microphone.

Under these conditions, the instantaneous pressure p at the microphone may be written[8] as

$$p = \frac{A}{x} \cos k(ct - x). \tag{1}$$

Although (1) pertains to a point source of sound for the theoretical analysis, it must be emphasized that, in practice, the actual microphone is used close to the mouth and the effects of mouth-cavity resonances, directivity of source, and body reflections will be present, causing the sound wave to deviate from a true spherical wave. Differentiating (1) with respect to x, there results

$$\frac{\partial p}{\partial x} = \frac{Ak}{x} \sin k(ct - x) - \frac{A}{x^2} \cos k(ct - x). \tag{2}$$

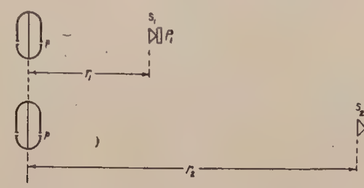


Fig. 2—Relative source and microphone locations used in calculating discriminations.

Obtaining the root-mean-square value of the pressure variation, we have

$$\left[\frac{\partial p}{\partial x}\right]^2 = \frac{A^2k^2}{x^2} \sin^2 k(ct - x) - 2\frac{A^2k}{x^3} \sin k(ct - x) \cos k(ct - x) + \frac{A^2}{x^4} \cos^2 k(ct - x).$$

Now averaging over a half cycle we have since

$$\frac{1}{\pi}\int_0^\pi \sin^2 \theta d\theta = \tfrac{1}{2},$$

$$\frac{1}{\pi}\int_0^\pi \cos^2 \theta d\theta = \tfrac{1}{2}$$

$$\frac{1}{\pi}\int_0^\pi \sin\theta \cos\theta d\theta = 0$$

and substituting $\theta = k(ct - x)$ and $d\theta = kcdt$

$$\left[\frac{\partial p}{\partial x}\right]^2 = \frac{A^2k^2}{x^2}\frac{1}{2kc} + \frac{A^2}{x^4 2kc}$$

so

$$\left[\frac{\partial p}{\partial x}\right]_{\text{r-m-s}} = \frac{Ak}{x\sqrt{2kc}}\sqrt{1 + \frac{1}{k^2x^2}}. \tag{3}$$

If, however, the root-mean-square value is found from (3) under the condition that x is large so that $1/x^2$ may be neglected, we see that

$$\left[\frac{\partial p}{\partial x}\right]_{\substack{\text{r-m-s} \\ x \text{ large}}} = \frac{Ak}{x\sqrt{2kc}} \tag{4}$$

the ratio of (3) to (4) is then the ratio of the action of the microphone under the two conditions and is thus the discrimination against the more distant source.

$$\frac{\left[\frac{\partial p}{\partial x}\right]_{\substack{\text{r-m-s} \\ x=r_1}}}{\left[\frac{\partial p}{\partial x}\right]_{\substack{\text{r-m-s} \\ x=r_2}}} = \frac{\frac{P_1k}{r_1\sqrt{2kc}}\sqrt{1 + \frac{1}{k^2r_1^2}}}{\frac{P_2k}{r_2\sqrt{2kc}}} = \frac{P_1r_2}{P_2r_1}\sqrt{1 + \frac{1}{k^2r_1^2}}. \tag{5}$$

[8] H. F. Olson, "Elements of Acoustical Engineering," D. Van Nostrand and Co., New York, N. Y., 1939, p. 11.

Since, for equal amplitudes of pressure at the microphone,

$$\frac{P_1}{r_1} = \frac{P_2}{r_2} \tag{6}$$

the noise discrimination, which is given as a ratio of the absolute magnitudes, may be written as

$$\text{first-order average discrimination} = \sqrt{1 + \frac{1}{k^2 r_1^2}} \tag{7}$$

on the axis. Equation (7) predicts the average discrimination on the axis as a function of (kr_1). The average discrimination due to polar response characteristics or directional efficiency[9–11] is that given by (7) multiplied by the $\sqrt{3}$.

First-order average random discrimination

$$= \sqrt{3}\sqrt{1 + \frac{1}{k^2 r_1^2}} \quad \text{for point sources.} \tag{8}$$

This equation is plotted as a function of kr_1 in Fig. 10 (see Section IV) in which the ordinate is 20 $\log_{10}$ (average random discrimination).

III. Practical Microphone of the First Order

Equation (8) gives the discrimination of a first-order gradient microphone where the acoustic distance between the openings to the two sides of the diaphragm is infinitesimal, and consequently very small compared to the distances from the sound sources. In a practical microphone, the acoustic distance between the sound

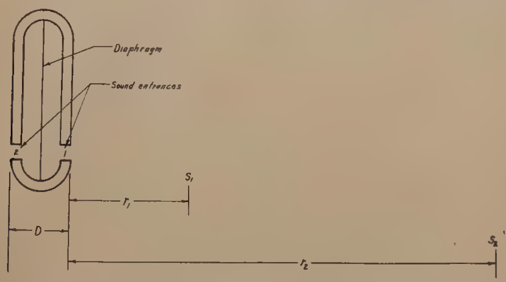


Fig. 3—Relative source and microphone locations used in calculation of discrimination of microphone with distance D between sound entrances.

entrances to the two sides of the diaphragm is comparable to the distance to the sound source. The effect of this finite spacing of the sound entrances is to decrease the discrimination from that predicted by (8) for distant sounds, and to limit the frequency range over which cancellation or discrimination is obtained. The smaller the spacing between the sound entrances, the higher the frequency at which discrimination is obtained and the better is the discrimination, because of smaller phase displacement.

The above statements are proved in the following manner. If the practical microphone has a spacing between sound entrances of D and is at distances of r_1 from the nearer sound source and r_2 from the distant sound source, the practical arrangement is as shown in Fig. 3.

[9] See p. 222 of footnote reference 8.
[10] B. B. Bauer, "Super-cardioid directional microphone," *Electronics*, vol. 15, p. 31; January, 1942.
[11] J. Weinberger, H. F. Olson, and F. Massa, *Jour. Acous. Soc. Amer.*, vol. 5, pp. 139–147; October, 1933.

Assuming, as for the theoretical case, that each source is producing equal pressure at the nearer sound entrance and that the wave form is spherical, then $p_1 = (P_1/r_1)\cos k(ct - r_1)$ is the pressure at the nearer sound entrance and $p_2 = (P_1/r_1+D)\cos k[ct-(r_1+D)]$ is the pressure at the second sound entrance when the nearer source S_1 is considered. Since D is comparable to r_1, there is a change in the magnitude of the pressure as well as a change in the phase of the pressure wave by an angular amount of kD. These results are shown in the vector diagram in Fig. 4.

Fig. 4—Vector representation of pressure actuating microphone diaphragm. Close point source.

Finding the resultant pressure p on the diaphragm

$$\Delta p = \sqrt{\left[\frac{P_1}{r_1} - \frac{P_1}{r_1+D}\cos kD\right]^2 + \left[\frac{P_1}{r_1+D}\sin kD\right]^2}$$

or

$$\Delta p = \frac{P_1}{r_1}\sqrt{1 + \left[\frac{r_1}{r_1+D}\right]^2 - \frac{2r_1}{r_1+D}\cos kD}. \tag{9}$$

By similar reasoning, the pressure at the two sound entrances due to the distant sound source is $p_1' = (P_2/r_2)\cos k(ct - r_2)$ and $p_2' = (P_2/r_2+D)\cos k[ct-(r_2+D)]$. In this case, however, since r_2 is very much larger than D, it may be assumed that $P_2/r_2+D = P_2/r_2$. There is, however, the same phase shift of kD, and the vector diagram becomes as shown in Fig. 5.

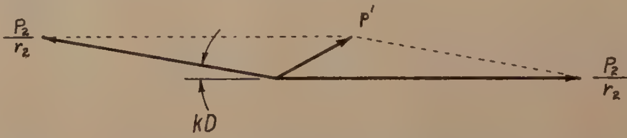


Fig. 5—Vector representation of pressure actuating microphone diaphragm. Distant point source.

Finding the resulting pressure p' on the diaphragm

$$\Delta p' = \sqrt{\left[\frac{P_2}{r_2} - \frac{P_2}{r_2}\cos kD\right]^2 + \left[\frac{P^2}{r_2}\sin kD\right]^2}$$

or

$$\Delta p' = \frac{P_2}{r_2}\sqrt{2 - 2\cos kD}. \tag{10}$$

The ratio of (9) to (10) then gives the discrimination

$$\frac{\Delta p}{\Delta p'} = \frac{\frac{P_1}{r_1}\sqrt{1 + \left[\frac{r_1}{r_1+D}\right]^2 - \frac{2r_1}{(r_1+D)}\cos kD}}{\frac{P_2}{r_2}\sqrt{2 - 2\cos kD}}. \tag{11}$$

However, by the conditions postulated for the problem $P_1/r_1 = P_2/r_2$ or equation (11) becomes

discrimination

$$= \sqrt{\frac{1 + \left[\frac{r}{r+D}\right]^2 - \frac{2r}{r+D}\cos kD}{2(1-\cos kD)}} \tag{12}$$

where D is the acoustic distance between the sound entrances and r the distance from the sound source to the nearest sound entrance Equation (12) gives the discrimination against sounds originating on the axis of the sound entrances. For random sounds, the discrimination is increased by about 5 decibels. Fig. 6 shows the effect of the sound-entrance spacing on the discrimination against random sounds for D equal to various fractions of r.

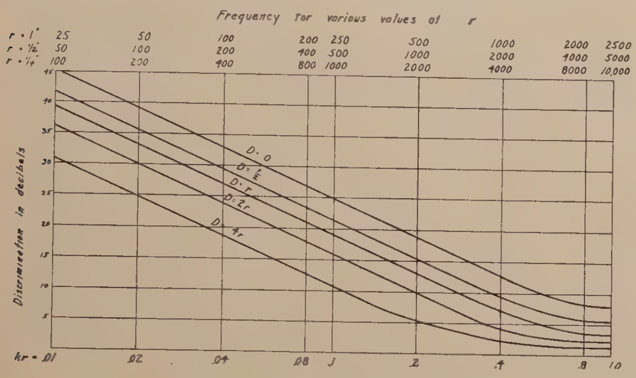


Fig. 6—Theoretical discrimination against random sounds for various spacings D of openings to front and back of diaphragm, and for various distances r from point source.

Equation (2) shows that the response of a gradient microphone is a function of the distance from the microphone to the source. From (3) it may be seen that the force available for actuating the diaphragm for sound waves of equal pressures is proportional to

$$\sqrt{k^2 + \frac{1}{r^2}}. \tag{13}$$

For very large values of r the force may be considered proportional to k, or in other words, the force is proportional to frequency. For low frequencies and small distances r, the force is proportional to $1/r$. The force available for actuating the diaphragm at various distances from the source versus frequency is plotted in Fig. 7, where the ordinate is 20 $\log_{10}$ (relative force). The frequency response of a microphone which is designed to have a flat response curve at a distance of one-quarter

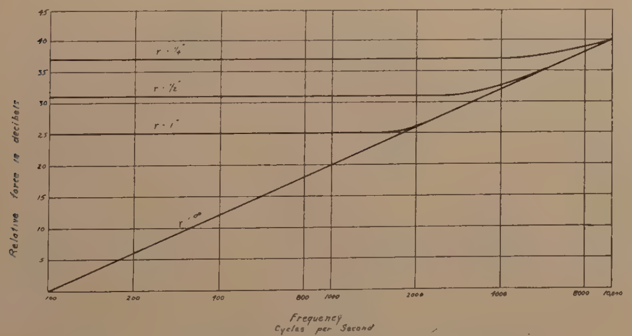


Fig. 7—Force available for actuating diaphragm.

inch from the source will no longer have a flat response curve at some other distance from the source. The frequency-response curves at various distances from the sound source, for a microphone designed to have a flat response curve at a distance of one-quarter inch, are shown in Fig. 8.

Fig. 9 shows a drawing of the T-45 lip microphone and its equivalent acoustic-mechanical circuit. Any unbalance in the impedances on the two sides of the microphone will cause the discrimination to suffer. The

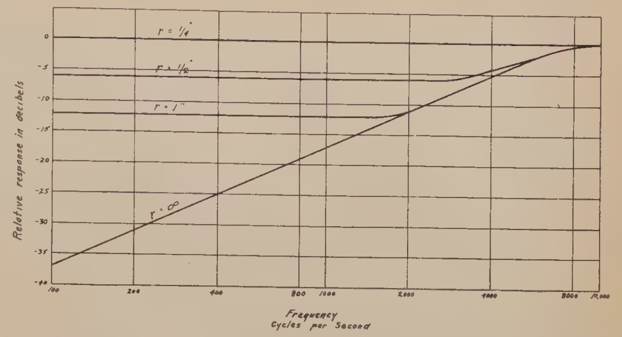


Fig. 8—Theoretical response at various distances r from source for microphones having a flat response $\frac{1}{4}$ inch from source.

head and mouth of the speaker will upset the impedance balance, causing the discrimination to suffer somewhat at the higher frequencies.[12]

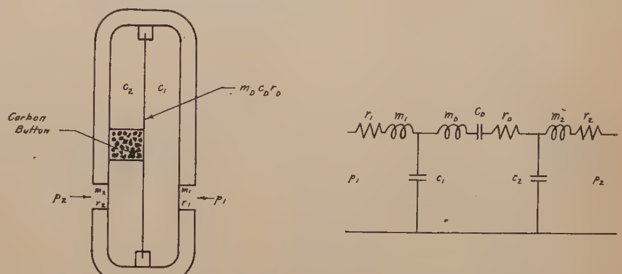


Fig. 9—Diagram of differential microphone and its equivalent acoustic-mechanical circuit.

IV. Theoretical Second- and Higher-Order Differential Microphones

It is of interest to see how the discrimination of the differential microphone changes as it changes from a first-order gradient to a second-, third-, fourth-, etc., order gradient microphone. In order to obtain the characteristics of a second-order differential, it is necessary to differentiate (2) again, which gives the following results:

$$\frac{\partial^2 p}{\partial x^2} = \frac{-k^2 A}{x}\cos k(ct-x) - \frac{2kA}{x^2}\sin k(ct-x) + \frac{2A}{x^3}\cos k(ct-x) \tag{14}$$

performing the averaging process on (14) as for the first-order differential case

$$\left[\frac{\partial^2 p}{\partial x^2}\right]_{\text{r-m-s}} = \frac{k^2 A}{x\sqrt{2kc}}\sqrt{1+\frac{4}{k^4x^4}}. \tag{15}$$

Again forming the ratio for the close to distant conditions

$$\frac{\left[\frac{\partial^2 p}{\partial x^2}\right]_{\substack{\text{r-m-s}\\ x=r_1}}}{\left[\frac{\partial^2 p}{\partial x^2}\right]_{\substack{\text{r-m-s}\\ x=r_2}}} = \frac{\frac{k^2P_1}{r_1\sqrt{2kc}}\sqrt{1+\frac{4}{k^4r_1^4}}}{\frac{k^2P_2}{r_2\sqrt{2kc}}} = \frac{P_1r_2}{P_2r_1}\sqrt{1+\frac{4}{k^4r_i^4}}.$$

[12] G. G. Muller, R. Black, and T. E. Davis, *Jour. Acous. Soc. Amer.*, vol. 10, pp. 6–14; July, 1938.

Substituting as previously for P_1 and using the average discrimination the following results:

$$\text{second-order average discrimination} = \sqrt{1+\frac{4}{k^4r_1^4}} \quad (16)$$

on the axis.

To find the average random discrimination, (16) must be corrected for the polar characteristic[9] which results in

second-order average random discrimination

$$= \sqrt{5}\sqrt{1+\frac{4}{k^4r_1^4}}. \quad (17)$$

The third-order differential is obtained similarly by differentiating (14) again resulting in the following equation:

$$\frac{\partial^3 p}{\partial x^3} = \frac{-k^3A}{x}\sin k(ct-x) + 3\frac{k^2A}{x^2}\cos k(ct-x)$$
$$+\frac{6Ak}{x^3}\sin k(ct-x) - \frac{6A}{x^4}\cos k(ct-x). \quad (18)$$

Performing the averaging process on (18) there results the following equation:

$$\left[\frac{\partial^3 p}{\partial x^3}\right]_{\text{r-m-s}} = \frac{k^3A}{x\sqrt{2kc}}\sqrt{1-\frac{3}{k^2x^2}+\frac{36}{k^6x^6}}. \quad (19)$$

Evaluating this for the boundary values assumed and forming the ratio for discrimination there results:

$$\frac{\left[\dfrac{\partial^3 p}{\partial x^3}\right]_{\substack{\text{r-m-s}\\ x=r_1}}}{\left[\dfrac{\partial^3 p}{\partial x^3}\right]_{\substack{\text{r-m-s}\\ x=r_2}}} = \frac{\dfrac{k^3P_1}{r_1\sqrt{2kc}}\sqrt{1-\dfrac{3}{k^2r_1^2}+\dfrac{36}{k^6r_1^6}}}{\dfrac{k^3P_2}{r_2\sqrt{2kc}}}$$
$$= \frac{P_1r_2}{P_2r_1}\sqrt{1-\frac{3}{k^2r_1^2}\ \frac{36}{k^6r_1^6}}.$$

Average discrimination of a third-order microphone becomes

third-order average discrimination

$$= \sqrt{1-\frac{3}{k^2r_1^2}+\frac{36}{k^6r_1^6}} \quad (20)$$

on the axis.

For random noise and effect of polar characteristic,[9] (20) becomes

third-order average random discrimination

$$= \sqrt{7}\sqrt{1-\frac{3}{k^2r_1^2}+\frac{36}{k^6r_1^6}}. \quad (21)$$

The fourth, fifth, and higher orders are obtained in a similar manner, giving the following results:

fourth-order average random discrimination

$$= \sqrt{9}\sqrt{1-\frac{8}{k^2r_1^2}+\frac{576}{k^8r_1^8}} \quad (22)$$

fifth-order average random discrimination

$$= \sqrt{11}\sqrt{1-\frac{15}{k^2r_1^2}+\frac{40}{k^4r_1^4}+\frac{14400}{k^{10}r_1^{10}}}. \quad (23)$$

Equations (17), (21), (22), and (23) are plotted in Fig. 10.

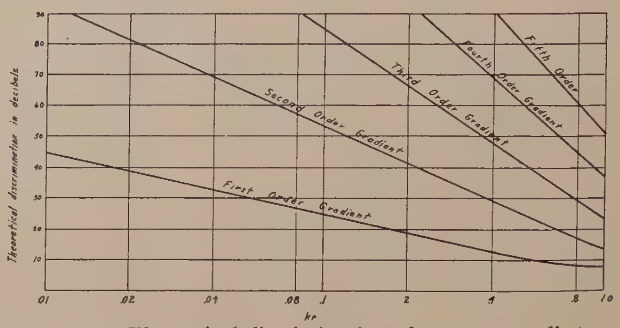


Fig. 10—Theoretical discrimination of pressure gradient microphones of various orders.

V. Theoretical nth-Order Differential Microphone

The discrimination of the nth-order differential microphone may be arrived at as follows: Using Leibnitz's rule for the nth derivative of a product,

$$\frac{\partial^n(\mu v)}{\partial x^n} = \frac{v\partial^n\mu}{\partial x^n} + n\frac{\partial v\partial^{n-1}\mu}{\partial x\partial x^{n-1}} + \frac{n(n-1)}{2!}\frac{\partial^2 v}{\partial x^2}\frac{\partial^{n-2}\mu}{\partial x^{n-2}}$$
$$+\frac{n(n-1)(n-2)}{3!}\frac{\partial^3 v}{\partial x^3}\frac{\partial^{n-3}\mu}{\partial x^{n-3}} + \cdots \frac{\mu\partial^n v}{\partial x^n} \quad (24)$$

where in the case at hand, since $p=(A/x)\cos k(ct-x)$, the function $\mu = x^{-1}$ and the function v is $\cos k(ct-x)$.

Using (24), it is possible to find the discrimination for any order differential microphone. It is of special interest to find the discrimination of the nth order. From (4), (15), and (19) it is seen that the nth gradient of pressure, when the source is a large distance away, is given in general by

$$\left[\frac{\partial^n p}{\partial x^n}\right]_{\substack{\text{r-m-s}\\ x=r_2}} = \frac{k^nP_2}{r_2\sqrt{2kc}}. \quad (25)$$

If the nth-order differential is obtained using (24), the following equation results:

$$\frac{1}{n!}\left[\frac{\partial^n p}{\partial x^n}\right] = \frac{A}{x^{n+1}}\left\{(-1)^n\cos k(ct-x)\left[1-\frac{(kx)^2}{n(n-1)}\right.\right.$$
$$\left.+\frac{(kx)^4}{n(n-1)(n-2)(n-3)}\cdots\right]$$
$$+(-1)^{n-1}\sin k(ct-x)\left[\frac{kx}{n}-\frac{(kx)^3}{n(n-1)(n-2)}\right.$$
$$\left.\left.+\frac{(kx)^5}{n(n-1)(n-2)(n-3)(n-4)}\cdots\right]\right\} \quad (26)$$

as n approaches infinity the coefficient of the cosine term in (26) approaches unity, while the coefficient of the sine term vanishes; hence, in the limit, (26) becomes

$$\frac{1}{n!}\left[\frac{\partial^n p}{\partial x^n}\right] = \frac{A(-1)^n}{x^{n+1}}\cos k(ct-x).$$

Performing the averaging operation on this expression and rewriting,

$$\left[\frac{\partial^n p}{\partial x^n}\right]_{\substack{\text{r-m-s}\\ x=r_1}} = \frac{n!P_1}{r_1^{(n+1)}}\left[\frac{1}{\sqrt{2kc}}\right]. \tag{27}$$

Dividing (27) by (25), the discrimination becomes

$$\frac{\left[\frac{\partial^n p}{\partial x^n}\right]_{\substack{\text{r-m-s}\\ x=r_1}}}{\left[\frac{\partial^n p}{\partial x^n}\right]_{\substack{\text{r-m-s}\\ x=r_2}}} = \frac{\frac{n!P_1}{r_1^{(n+1)}}\left[\frac{1}{\sqrt{2kc}}\right]}{\frac{k^nP_2}{r_2\sqrt{2kc}}} = \frac{n!P_1}{k^nP_2}\left[\frac{r_2}{r_1^{(n+1)}}\right]$$

or again finding the average discrimination

$$n\text{th-order average discrimination} = \frac{n!}{k^n r_1^n} \text{ on the axis.} \tag{28}$$

It is clear that as n becomes infinite the discrimination of an nth-order differential microphone becomes infinite.

As the order of the gradient becomes larger, the directional efficiency increases as the multiplying factor in (8), (17), (21), (22), and (23) show. The increased discrimination due to polar response may be written in general as $\sqrt{2n+1}$. As n approaches infinity, the increased noise discrimination due to polar response is increased by an infinite amount. This corresponds to a decreasing solid angle over which pickup of noise is obtained. The response of the microphone is now limited to that produced by sound sources lying on the axis of the sound entrances. Consequently the average random discrimination is increased over that predicted by (28).

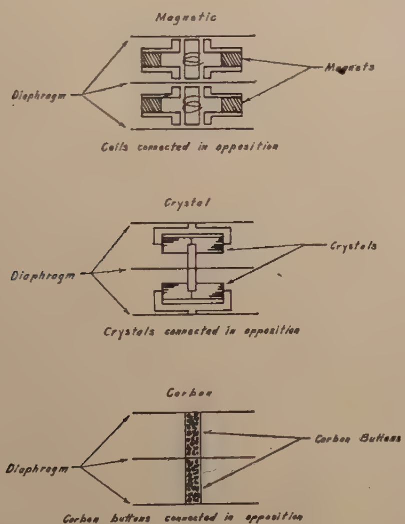


Fig. 11—Higher-order gradient microphones.

VI. Differential Microphones of a Higher Order

There are many ways in which a higher-order differential microphone may be designed. Two closely spaced first-order differential dynamic microphones with their voice coils connected in opposition will be a second-order differential microphone. Two of these combinations connected differentially will give a third-order differential microphone, etc. Diagrammatic representations of other types of higher-order differential microphones are shown in Fig. 11. There are certain practical limitations to the construction of higher-order differential microphones. Each first-order differential unit used in a higher-order differential microphone must have very nearly identical sensitivity and response. This problem of balancing limits the practical use of higher-order differential microphones. The signal-to-noise ratio of the higher-order series is, of course, considerably better than that for a first-order differential; but the first-order differential gives sufficient discrimination against noise for practically all applications. The response of a higher-order differential microphone also varies more with the distance from the source than does a first order.

The discrimination curves of higher-order differential microphones indicate that they would be much better for use, under severe feedback conditions, than a first order. This is not true in practice, however, as the mouth acting as a Helmholtz resonator upsets the balance causing the system to oscillate before the full discrimination, against unwanted sounds originating at a distance, can be realized.

VII. Conclusion

From the equations developed and the curves plotted for the differential microphone, irrespective of the order of pressure gradient which operates it, it is clearly seen that the differential microphone is ideally suited for noise cancellation in the usual noise fields. The frequencies encountered in the noise of aircraft, tanks, gun fire, trains, etc., are predominantly in the lower frequency range and the differential microphone produces more noise discrimination at the lower frequencies for a fixed position of the microphone.

It has been shown that the noise discrimination of differential microphones increases rapidly with the order of pressure gradient upon which they operate and that the nth-order gradient would give the most desirable results theoretically. In practice, however, the construction of higher-order gradient microphones presents difficulties due to balancing of the individual first-order gradient units used in the higher-order gradient series.

Discussion on

"Exalted-Carrier Amplitude- and Phase-Modulation Reception"*

Murray G. Crosby

A. H. Taylor:[1] I here read with particular interest the paper by Crosby on "Exalted Carrier Reception."* When I was a student with Professor Barkhausen at the Sächsische Technische Hochschule, Dresden, in 1937–38, Professor Barkhausen's chief assistant, H. Oltze, demonstrated a broadcast receiver embodying the same principles of carrier reinforcement. Herr Oltze used one mixer only, with two separate intermediate-frequency amplifiers at the same intermediate frequency, for program and supplementary carrier. There was no automatic frequency control. Automatic frequency control was not necessary because there was no crystal filter or limiter; modulation was eliminated from the carrier channel by sharp inductance-capacitance tuning plus a fast automatic volume control having a time constant of 1/20,000 second. Like Mr. Crosby, Herr Oltze fed his supplementary carrier to the second (recombining) detector via a phase-adjusting network; however, the relative phases of normal and supplementary carriers seemed unimportant provided the latter was large enough.

The receiver was demonstrated by tuning in an out-of-town station (Praha, Č.S.R.) and then setting beside the receiver a laboratory oscillator tuned to give a loud audio whistle with the distant transmitter, with no carrier reinforcement. When the reinforcement was brought up, the whistle all but vanished and the program could be enjoyed. This seems to disagree with the experience of Mr. Crosby, who states that intercarrier heterodyne notes are not reduced by carrier reinforcement.

Herr Oltze did this work in his spare time, more or less as a stunt. He did not know of the work of Polkinghorn and Schlaack, and asked me whether I though he could get a U. S. Patent. I think he should get whatever credit is due him, particularly since he was a perfect gentleman and no Nazi.

Murray G. Crosby:[2] Mr. Taylor's interesting letter has brought to the fore a possible misunderstanding that I am happy to have this opportunity to clarify.

I have reference to the apparent disagreement as to the reduction of intercarrier heterodyne. As I see it, there is no disagreement as to results. I believe the results described by Mr. Taylor would be duplicated on the exalted-carrier receivers with which I have worked.

* Murray G. Crosby, "Exalted-Carrier Amplitude- and Phase-Modulation Reception," Proc. I.R.E., vol. 33, pp. 581–591; September, 1945.
[1] Naval Research Laboratory, Washington 20, D. C.
[2] Paul Godley Company, Upper Montclair, N. J.

The apparent disagreement is brought about by a difference in interpretation of results, which in turn, was brought about by a difference in test conditions.

The statement that intercarrier heterodynes were not suppressed was based on the use of carriers which were both modulated. Under these test conditions, the introduction of carrier exaltation prevents the modulation from the stronger, undesired carrier from masking the modulation on the weaker, desired carrier. Switching to the exalted-carrier condition then coverts the receiver output from undesired to desired modulation, without change in the amplitude of the intercarrier whistle.

In the conditions of Herr Oltze's experiments, the strong, undesired carrier was unmodulated. In the absence of exaltation this condition would leave a predominance of intercarrier heterodyne, with weak, desired modulation in the background. The introduction of carrier exaltation would bring up the level of the desired modulation, but the level of the whistle would remain the same. Where the confusion in interpreting the results arises is in the fact that the output amplitude of the heterodyne remains constant while the desired-modulation output is increased by the exaltation. Perhaps a more accurate and descriptive term is "desired-modulation-to-whistle ratio." We would both agree that this ratio is improved by the use of carrier exaltation, regardless of whether the undesired carrier is modulated or not. Such a term eliminates the necessity of having to choose between the two test conditions which involve absence, or presence, of modulation on the interfering carrier.

With regard to credit to Herr Oltze, I would not be willing to concede priority on the basis of the dates mentioned in Mr. Taylor's letter. Receivers of the type mentioned were demonstrated at the Riverhead RCA Laboratories considerably earlier than 1937. In this respect reference is made to my U. S. Patent No. 2,065,565 filed June 13, 1932, and patented December 29, 1936, which describes a working model of a receiver of this type.

A. H. Taylor: Mr. Crosby's reply to my comments answers all questions which I had in mind.

This is apparently another of the many instances where the first inventor of a device does not immediately publish, and subsequently someone else invents the same thing quite independently, without knowing of the original inventor's work.

Correspondence

Correspondence on both technical and nontechnical subjects from readers of the PROCEEDINGS OF THE I.R.E. and WAVES AND ELECTRONS is invited, subject to the following conditions: All rights are reserved by the Institute. Statements in letters are expressly understood to be the individual opinion of the writer, and endorsement or recognition by the I.R.E. is not implied by publication. All letters are to be submitted as typewritten, double-spaced, original copies. Any illustrations are to be submitted as inked drawings. Captions are to be supplied for all illustrations.

"Historic First" Electronic Lecture

To the Editor:

In the February, 1945, issue of the PROCEEDINGS, in the editorial captioned "Electronic Papers," Past President H. M. Turner reminds us that The Institute of Radio Engineers has been actively interested in electronics from the early days.

This is indeed true, as witness the first public demonstration of the vacuum tube as an amplifier, the first such in this country at least, made before the Institute in the late fall of 1913 by Dr. Lee de Forest himself.

The report of this "historic first" electronic lecture is published as a paper on pages 15–30 of the March, 1914, issue of the PROCEEDINGS. I believe it was you, Mr. Editor, in your youthful enterprise, who drafted that paper after the lecture and then had it passed upon by de Forest. For as I remember it, upon this occasion de Forest with his usual enthusiasm launched into the subject, talked and demonstrated to his heart's content, without confining himself to a text. Incidentally I think the Editor is to be congratulated upon his enterprise and reporting ability, whereby we do have an appropriate record of this important meeting. For certainly "Doc" himself was too impatient to have prepared the paper and probably would not have remembered just what he did say! Incidentally the *discussion* of the paper as published is especially interesting, touching as it did on feedback and what not.

It happened that the present writer and a boyhood amateur pal of his with whom he is still associated here in the Bell Telephone Laboratories, namely Austen M. Curtis, were among those who attended this meeting. I wonder how many other old-timers recall it. Fortunately what de Forest disclosed and said so impressed me at the time that I sent a memorandum on the meeting to my departmental head who was then Elam Miller, transmission and protection engineer of the American Telephone and Telegraph Company. Since this little report brings out some additional aspects of the meeting, as seen by one who was then a sort of hybrid radio-wire-telephone engineer, a verbatim copy of the memorandum is attached.

Incidentally, de Forest's paper, as published in the PROCEEDINGS, is given in a footnote as having been delivered on December 3, 1913. The date appears to be in error. The memorandum which I wrote on the day following the meeting is dated November 6. A second memorandum which I wrote on the same day, on the subject of the vacuum tube used as a power limiting device, helps to confirm the date of de Forest's lecture as that of November 5. The *New York Herald* for Friday, November 7, 1913, carried a little news item of de Forest's lecture, pertaining to the possible use of the vacuum tube on the then projected transcontinental telephone line, and referred to the meeting as having been held at Columbia University "last Tuesday night." This would have been November 4. Probably the meeting was on the usual Wednesday night, which would make the date November 5.

Sincerely,

LLOYD ESPENSCHIED
Bell Telephone Laboratories
New York, N. Y.

The Audion as an Amplifier—Paper by Dr. Lee de Forest.

November 6, 1913.

Memorandum for Mr. Miller,
Transmission and Protection Engineer.

At a meeting of the Institute of Radio Engineers held at Columbia University November 5, Dr. Lee de Forest gave a paper accompanied by demonstrations on the Audion as ~~an~~ amplifying device. Following are the principal points of interest which were discussed.

The use of the Audion as a detector and as an amplifier in radio communication was first considered. Emphasis was laid upon the distinct difference between the Audion and Fleming's vacuum rectifier detector; that there is a difference is made evident by the fact that the maximum possible converter efficiency of the Fleming valve is 50 per cent, while that of the Audion may easily be several hundred per cent. As evidence of the sensibility of the Audion in radio receiving, mention was made of instances of receiving long-distance messages. It was claimed that signals which were practically inaudible when using the Audion simply as a detector may be converted into signals of commercial volume by means of the Audion amplifier. He stated that the Audion was used as an amplifier in the San Francisco-Los Angeles Poulsen wireless transmission. The signals were transmitted at about 60 words per minute and at the receiving end were amplified by means of the Audion and recorded on the telegraphone. The telegraphone, running at a low speed, then repeated the signals to a receiving operator who transcribed them on a typewriter.

Referring to the use of an Audion as a telephone amplifier mention was made of the difficulty met in obtaining lines sufficiently well balanced to be efficiently used in connection with a repeater. Proper circuits are now being worked out and it was thought probable that the Audion would be very successfully applied in telephone transmission. A demonstration of the amplification of the telephone currents was made using an ordinary telephone receiver as a magneto transmitter and three Audions connected in tandem as the receiving amplifier. Speech was transmitted from an adjacent room, amplified by means of the Audion, and transmitted to the audience through loudspeaking receivers. In a similar way, wireless signals which had previously been recorded on the telegraphone were made audible to the entire audience. In connection with the use of the Audion in telephone transmission, Dr. de Forest mentioned a possible telephone system organized as follows:

Use a magneto transmitter at the subscriber's station instead of the microphone transmitter for the purpose of more faithfully reproducing the voice.

Transmit the currents so generated (without any battery) to the central office, possibly over subscriber loop conductors much smaller than those now used.

At the central office pass the transmitted currents through an Audion amplifier which directly feeds the trunk.

At the distant central office possibly again amplify the telephone currents by means of an Audion and pass them to the subscriber loop and receive in the ordinary way.

A discussion was taken part in by Mr. John Stone Stone. He considered the Audion the most promising device he knew of for detecting radio signals and for amplification purposes. He recalled that the most objectionable feature to a system of high-frequency wire transmission, such as has recently been brought forward by Major Squier, is the very high attenuation. He thought it quite probable that this objectionable feature may be entirely eliminated by the use of the Audion amplifier and that high-frequency telephone transmission may thus be placed on an economical basis at least comparable with that of the present system of telephone transmission. He pointed out that high-frequency transmission was practically void of distortion and that the multiplexing of such a line, for either telephone or telegraph purposes, is a valuable feature.

(Initialed) L.E.

NEW YORK HERALD
FRIDAY, NOVEMBER 7, 1913

(Article appearing on the Editorial Page, page 12)

MAY TALK TO THE PACIFIC
Telephone Service Across Continent
Believed Near By The Use Of
Audion Amplifier

That direct telephone service between this city and San Francisco will begin soon was the assertion of many who attended the monthly meeting of the Institute of Radio Engineers at Columbia University last Tuesday night.

Dr. Lee de Forest, inventor of the new Audion amplifier, or relay, demonstrated the possibilities of his invention, and declared that through the amplification of minute electric currents results will be accomplished heretofore considered impossible.

Representatives of the United States Navy who were present showed much interest in the invention as a possible aid in wireless operation.

A true transcription made today in the Old-Newspaper Room of the New York Library, 137 West 25 Street, New York City, by Lloyd Espenschied, July 6, 1944.

Phase-Inverter Circuit

A brief description of a so-called "phase-inverter" circuit written by D. L. Drukey from a very famous locale, recently appeared in the the PROCEEDINGS.[1] An interesting application of this circuit was used in a radio receiver designed for the Northern Electric Company, Limited, in 1936.

A schematic diagram of the pertinent part of the circuit is shown in Fig. 1. A type 6C5 tube was used to drive a pair of type 6V6G tubes connected in push-pull, with circuit-component values as shown. It is seen that the grid of one output tube was driven from a 10,000-ohm resistor in the plate circuit of the type 6C5 tube, and the other output-tube grid was driven from a 10,000-ohm resistor in the cathode circuit. Grid bias for the driver tube was secured from an additional unbypassed resistor in its cathode circuit. While this resulted in some loss of gain, it was undesirable for standardization reasons to secure bias from a division of the 10,000-ohm cathode resistor. As the type 6C5 tube followed a high-impedance stage it was necessary to introduce a resistance-capacitance filter to eliminate alternating voltage from the cathode coupling resistor from the grid circuit of the type 6C5 tube. Under the circuit conditions shown the type 6C5 tube operates with the following constants: plate current, 2.7 milliamperes; amplification factor, 17.5; plate resistance 10,300 ohms; mutual conductance, 1700 micromhos; undistorted output, 40 volts peak, grid-to-grid.

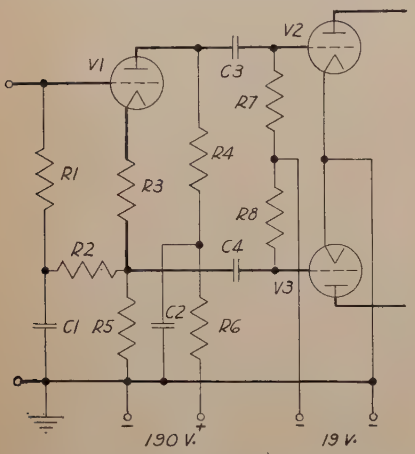


Fig. 1—Detailed circuit of unbalanced-to-balanced impedance conversion stage. Circuit values as follows:

Microfarads	Ohms	Tube
$C1=0.05$	$R1=500,000$	$V1=$ 6C5
$C2=8$	$R2=500,000$	$V2=$ 6V6G
$C3=0.02$	$R3=1500$	$V3=$ 6V6G
$C4=0.02$	$R4=10,000$	
	$R5=10,000$	
	$R6=5000$	
	$R7=100,000$	
	$R8=100,000$	

The stage gain is 3 decibels and there is 17 decibels of negative feedback. This circuit therefore not only achieves the high degree of balance and the independence of tube characteristics mentioned by Drukey, but also suppresses by a large factor any hum, tube noise, or distortion generated in the driver stage. It also results in an equivalent alternating-current impedance across the grids of the push-pull tubes of only about 3200 ohms, which is of great practical importance with many types of tubes. If *R3* is adequately bypassed or otherwise removed from the feedback path, the gain of the stage increases to 5 decibels and the grid-to-grid impedance is reduced to less than 1000 ohms.

C. B. FISHER,
F. T. Fisher's Sons, Ltd.,
Montreal, Que., Canada

My thanks to Mr. C. B. Fisher for his comment on my letter on the phase inverter. This was not presented as an original idea, although to my knowledge it was, but rather as a circuit which warranted further use in the electronic art.

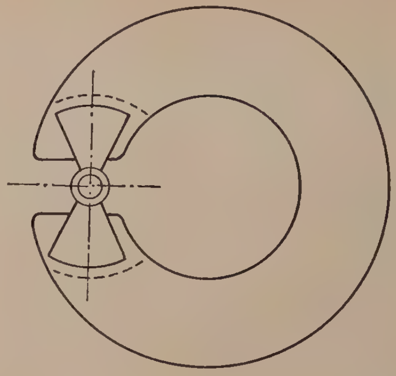
D. L. DRUKEY
17 Outer Dr.
Oak Ridge, Tenn.

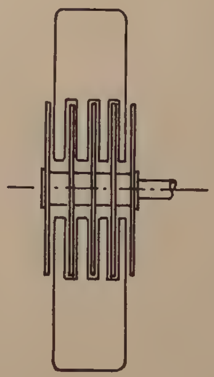
[1] PROC. I.R.E., vol. 33, p. 722; October, 1945.

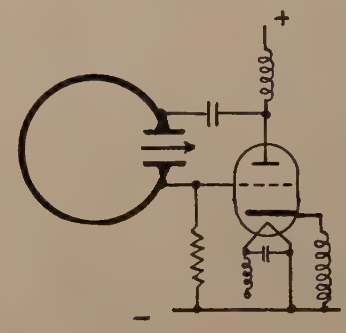
Asymmetrical Butterfly Circuit

I have read the excellent paper by Mr. E. Karplus on "wide-range tuned circuit and oscillators for high frequencies" in the July issue of the PROCEEDINGS of the I.R.E. It may be of interest to your readers to know that late in 1941, while working in the laboratories of the Electric and Musical Industries, Ltd., Hayes, England, on a radar equipment, I developed an asymmetrical butterfly circuit which performed exceedingly well. It is described in detail in the British Patent No. 563,468. A considerable amount of work on lumped tuned circuits has been done by W. S. Percival of the abovementioned laboratories, of whose ideas my invention was a further development.

The mechanical build-up and schematic diagram are shown in the accompanying illustrations. The tube used was a Mullard RL.16 and the frequency range of the oscillator circuit was restricted deliberately







to cover 290 to 350 megacycles per second. The stability of the circuit proved to be excellent.

A. LANDMAN
2 Monkswood Gardens
Ilford, Essex
England

Contributors

Carl Beckman

Carl Beckman (A'42) was born on August 27, 1915, in Philadelphia, Pennsylvania. He received the B.A. degree in physics from Temple University in 1937, the B.S. degree in electrical engineering from the University of Pennsylvania in 1940, and the M. S. degree in electrical engineering from Stevens Institute of Technology in 1943. Mr. Beckman has been an electronics engineer with the Westinghouse Electric Corporation, Bloomfield, New Jersey, since 1940. He is a member of Eta Kappa Nu, the American Institute of Electrical Engineers, the American Physical Society, the Franklin Institute, and the Montclair Society of Engineers.

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D. Rogers Crosby (A'36) was born on October 12, 1911, at Lockport, New York. He received the E.E. degree from Rensselaer Polytechnic Institute in 1934, and the M.S. degree in communication engineering from Harvard University, in 1935.

❖

Gordon L. Fredendall

From 1935 to 1941, Mr. Crosby was employed by the Federal Telephone and Telegraph Company, at Newark, New Jersey, engaged in experimental work on radio printer circuits and very-high-frequency link circuits for control of communication transmitters. Since 1941 he has been associated with the Radio Corporation of America at Camden, New Jersey. He is engaged in development work on problems associated with high-powered transmitters, particularly methods and means for measuring high voltage and high power.

Mr. Crosby is an Associate of the American Institute of Electrical Engineers, and a member of Sigma Xi.

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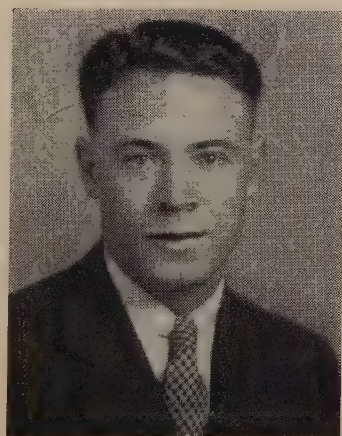
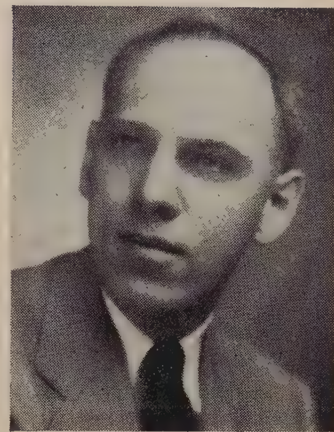


Harold E. Ellithorn

Harold E. Ellithorn (A'36–M'44) was born on October 11, 1911, at Detroit, Michigan. He received the B.S. degree in electrical engineering from Union College in 1934, and the M.S. degree in communication engineering from Harvard University in 1935.

From 1935 to 1936 Mr. Ellithorn was employed in the engineering department of Sylvania Corporation, in Salem, Massachusetts, and in 1936 he became supervisor of the engineering laboratory of the radio-tube division. From 1938 to 1940 he was a graduate student in physics at the University of Notre Dame, and in 1940 he was appointed an instructor in electrical engineering at that institution. He became an assistant professor in 1943, a post which he has held to date, and since 1944 Mr. Ellithorn has also served as special project engineer and consultant for the Electro-Voice, Inc., in South Bend, Indiana.

He is a member of Eta Kappa Nu and the Acoustical Society of America, and an Associate Member of Sigma Xi and the American Institute of Electrical Engineering.



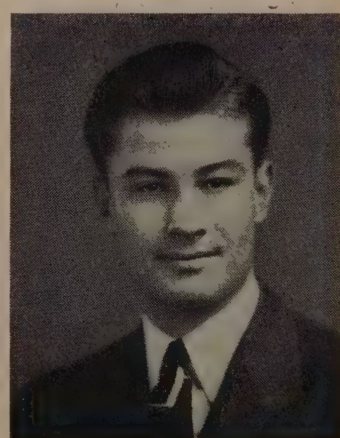
D. Rogers Crosby

❖

Gordon L. Fredendall (A'41) received the Ph.D. degree from the University of Wisconsin. From 1931 to 1936 he taught electrical engineering and mathematics, and engaged in research work in mercury-arc phenomena at the University of Wisconsin. Since 1936 he has been with the Radio Corporation of America, working on television research. He is at present located in Princeton, New Jersey, at the RCA Laboratories.

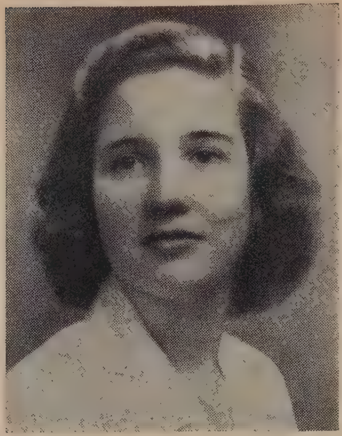
❖

C. Herbert Gleason (A'42) was born at Halstead, Kansas, on October 1, 1918. He received the B.S. degree in electrical engineering from the University of Kansas in 1940, and the M.S. degree in electrical engineering from the University of Missouri in 1942. He was a teaching assistant at the University of Missouri during 1940 and 1941 and an instructor at that University in 1941 and 1942. Since 1942 Mr. Gleason has been with the electronics engineering division of the Westinghouse Electric Corporation, Bloomfield, New Jersey. He is a member of Sigma Xi and Tau Beta Pi.



C. Herbert Gleason

A. C. Schroeder



Carol H. Pennypacker

Myron S. Wheeler

A. C. Schroeder (A'38) was born at West New Brighton, Staten Island, N. Y., on February 28, 1915. He received the B.S. degree in electrical engineering from the Massachusetts Institute of Technology in 1937, and the M.S. degree from the same institution in the same year. He joined the Radio Corporation of America in 1937, and is now engaged in television research at the RCA Laboratories in Princeton, New Jersey. He is a member of the American Association for the Advancement of Science.

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Kurt Schlesinger (A'41) was born on April 20, 1906, at Berlin, Germany. In 1928 he received the engineer's diploma, and in 1929 the degree of Doctor of Applied Physics, both from Technische Hochschule in Berlin. From 1929 to 1930 he was research physicist in Ardenne Research Laboratory, Berlin, and from 1931 to 1937 was chief engineer in the television department of Loewe Radio Company, also in Berlin. In 1938 Dr. Schlesinger became affiliated with the Radio and Cables-Grammont in Paris, France, where he devoted his time to television development. From 1941 to 1944 he was a research engineer for the Radio Corporation of America, attached to the laboratory at Purdue University. He is now associated with the Columbia Broadcasting System in New York City.

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Carol H. Pennypacker was born on March 1, 1922, at Haddonfield, New Jersey. She received the A.B. degree in mathematics from Smith College in 1943, and, since graduation, has been employed in a research division of the Radio Corporation of America, at Camden, New Jersey.

❖

D. L. Waidelich (S'37–A'39–SM'44) was born on May 3, 1915, at Allentown, Pa. He received the B.S. degree in electrical engineering in 1936, and the M.S. degree in 1938, from Lehigh University. He was teaching assistant at Lehigh University from 1936 to 1938, and instructor and assistant professor of electrical engineering at the University of Missouri from 1938 to 1944. In the summer of 1937 he worked with Bell Laboratories, Inc., in the field of electronics research and in the summer of 1941 was a member of the electronic research staff of the Westinghouse Electric and Manufacturing Company, at Bloomfield, New Jersey.

During 1944 and 1945, Mr. Waidelich was with the Naval Ordnance Laboratory in Washington, D. C., as an electrical engineer and has now returned to the University of Missouri. He is a member of the American Institute of Electrical Engineers, the Society for the Promotion of Engineering Education, Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

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Myron S. Wheeler (A'43) was born at Pittsburgh, Pennsylvania, on January 15, 1920. He received the B.S. degree in electrical engineering in 1942 from the Pennsylvania State College. Since that time he has been doing graduate work at Stevens Institute of Technology and working with the Westinghouse Electric Corporation in a development section on radar. He is a member of Eta Kappa Nu and Sigma Tau.

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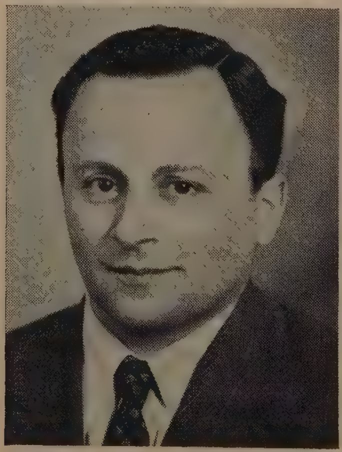
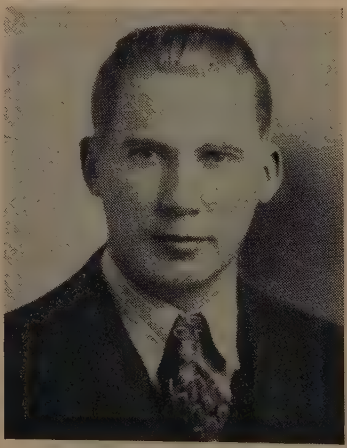
A. M. Wiggins (A'42) was born on January 8, 1911, at Plainview, Texas. He received the B.S. degree in electrical engineering from Texas Technological College in 1933, and the M.S. degree in electrical engineering in 1936, from the University of Texas. From 1936 to 1941 he was employed by Seismic Explorations, Inc., of Houston, Texas. In 1941 he became associated with the RCA Manufacturing Company, in Camden, New Jersey, working in the research department until 1942, when he was transferred to RCA Laboratories, in Princeton, New Jersey, doing acoustical research. From 1944 to the present, Mr. Wiggins has been a research engineer for the Electro-Voice, Inc., South Bend, Indiana.

He is a member of Sigma Xi and the Acoustical Society of America.

D. L. Waidelich



Kurt Schlesinger



A. M. Wiggins

WAVES AND ELECTRONS

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Institute News and Radio Notes

Books

PROCEEDINGS contents on page 47 P.

Donald B. Sinclair

Board of Directors—1946

Donald B. Sinclair was born at Winnipeg, Manitoba, Canada, on May 23, 1910. He attended the University of Manitoba from 1926 to 1929, and from 1929 to 1932 was enrolled in a co-operative course in electrical engineering at the Massachusetts Institute of Technology, from which he received the S.B. degree in 1931, the S.M. degree in 1932, and the Sc.D. degree in 1935.

During 1930 and 1931, Dr. Sinclair was associated with the Bell Telephone Laboratories and the New York Telephone Company, in New York City, and with the Western Electric Company, in Hawthorne, New Jersey. As a research assistant, from 1932 to 1935, and research associate during 1935 and 1936, he worked with the Massachusetts Institute of Technology on the electrocardiograph, electromechanical computing machines (differential analyzer and cinema integraph), and high-frequency measurements. Under co-operative arrangement with the General Radio Company, his doctorate research work on high-frequency impedance measurements was sponsored by that organization and done in its laboratories during 1934 and 1935.

In 1936 Dr. Sinclair became an engineer with the General Radio Company, and was appointed chief engineer in 1944. His work in this connection has dealt with general development and design of measuring instruments, with emphasis on high-frequency impedance-measuring equipment.

Dr. Sinclair was a radio amateur in Winnipeg from 1926 to 1929, under call VE4FV, and from 1928 to 1929 he was a part-time radio operator for the Western Canada Airways. His wartime activities include work as research associate and consultant with the radio research laboratory of Harvard University, and as a section member on the National Defense Research Committee. He served in North Africa at the Allied Force Headquarters in 1943, as technical observer with the United States Army Airforces.

Dr. Sinclair became a Junior member of The Institute of Radio Engineers in 1930, an Associate in 1933, a Member in 1938, and a Senior Member in 1943. He was awarded the grade of Fellow in 1944, "for the development and application of various types of networks for high-frequency measurement of impedance." He was elected a member of the Board of Directors of the Institute in 1945, and is active on the Papers, Papers Procurement, Television, and the Antenna Committees. He also served on the Transmitters and Antennas Committee from 1941 to 1944.

The New York Section of the Institute held its fourth Radio Pioneers' Dinner on November 8, 1945. Two of the honored guests, Rear Admiral Joseph R. Redman, U. S. Navy, Chief of Naval Communications, and Major General H. C. Ingles, Chief Signal Officer, War Department, addressed to Louis G. Pacent, Chairman of the General Committee for the dinner, the tributes to all radio and electronic engineers and workers which are presented below.

Army and Navy Letters of Recognition

On the occasion of your Radio Pioneers' Party, I welcome the opportunity to express the gratitude of the Signal Corps to the many radio-and-electronic engineers whose notable contributions to victory are now part of an imperishable historical record.

Demands, undreamed of by the founders of radio, were made by the Signal Corps upon those engineers and upon the entire electronics industry.

These demands were met even though it was necessary to create what virtually amounted to a new industry. Organizations and individuals placed the war needs of the nation above all other considerations and in the quality and quantity of their production and by their unselfish public service earned the acclaim of their country.

It is the earnest desire and confident hope of the Signal Corps that this close relationship and co-operation will not now be relaxed but will continue long into the future in order that our Army may be assured of unparalleled communications equipment.

Sincerely yours,
H. C. Ingles
Major General
Chief Signal Officer

Whenever a group of old-timers meet at a gathering such as this Radio Pioneers' Party it is customary to think back over the years and to reminisce about the progress that has been made within our field. It is doubtful that the most imaginative old-timer could predict the remarkable development which was to take place in radio. Both World Wars I and II proved to be a tremendous stimulus to the radio industry. The great strides we have made in electronics during the past few years are now beginning to be made known to the public. The result will be that the electronic industry will step up in importance to one of the front-rank fields of our daily life and will find outlets for its devices in every branch of human endeavor.

Our national defense will depend to a still greater extent upon the electronic devices in the near future. It is an important element in all our planning. Our position in world affairs is dependent upon the progress which we have made and hope to make in the future. The progress we make will be the result of the brain power and human efforts which lie behind all development work. The past history is remarkable but we may be certain that the future will be even more so. The Navy appreciates the good work which has been done and will co-operate in every way in order to assist in making future developments as fantastic to the present-day conceptions as is the present state compared with the crude wireless you pioneers experimented with in the early days.

The Navy extends its congratulations to the Pioneers who laid the ground work for the achievements of the present day. 73.

Joseph R. Redman
Rear Admiral, U.S. Navy
Chief of Naval Communications

The Historical Background of the Canadian Council of The Institute of Radio Engineers*

RALPH A. HACKBUSCH†, FELLOW, I.R.E.

WHILE the actual inception of the Canadian Council of The Institute of Radio Engineers is relatively recent, the events leading up to the formation of the Council date back to the year 1925.

During that year, a group of radio engineers in Toronto decided that they should set up a group representative of the radio engineers in Canada, and after mature discussion, petitioned the Board of Directors of The Institute of Radio Engineers for a charter setting up a Canadian Section on October 2, 1925.

This charter was granted on July 10, 1926, and the Canadian Section, as such, made considerable progress, partially due to the energetic leadership of the various executive committees, partially to the fact that Toronto became the center of the bulk of receiving-set manufacture in Canada prior to 1931, and partially to the fact that the University of Toronto offered a communications option in the electrical-engineering course of the School of Practical Science, dating back to 1922.

The fact that the Dean of the department of electrical engineering, several of the professors, and many of the lecturers at the university were interested in the work of the Institute, and the fact that the meetings have always taken place in the Electrical Building of the university, have contributed to the progress of the Institute in the Toronto area.

In August, 1930, due to the energetic work of the Toronto group, the Fifth Annual and First International Convention of the Institute took place in Toronto, during which convention joint meetings with the engineering committee of the Radio Manufacturers Association were featured, following a pattern instituted in the United States of America at the Radio Manufacturers Association Annual Meeting at Atlantic City the previous year.

The Canadian Section co-operated with the Rochester, Buffalo, Niagara Falls, and Cleveland Sections in establishing the successful I.R.E. Eastern Great Lakes District Convention, held in Rochester in the fall of 1929, which was the forerunner of the now annual Rochester Fall Meeting of members of the I.R.E. and of the Radio Manufacturers Association engineering department.

* Decimal classification: R060. Original manuscript received by the Institute, October 3, 1945.
† Stromberg-Carlson Company, Toronto, Ontario, Canada.

After the depression of 1930, many of the receiving-set manufacturers in the Toronto Area ceased to function, and many of the engineers who had been charter members of the Canadian Section of The Institute of Radio Engineers moved to Montreal, which was the center of the commercial activities of three large companies.

It was natural that this group should petition the Institute's Board of Directors for a charter for a Montreal section. This charter was granted on December 2, 1936, and formally presented to the Section on January 20, 1937.

Due to the close co-ordination of the members of the Montreal Section and the Toronto Section through close association in the activities of the Radio Manufacturers Association of Canada engineering committee over a period of years, the Canadian members of the Institute were kept informed of the activities of the two Sections.

With the advent of the war, Canada having declared war on Germany in September, 1939, there started a displacement of personnel from former centers and an increased interest in the field of radio and electronics. Naturally, there followed an increase in the Canadian membership of the Institute and a specific localization of interest in those areas where the war effort had established centers of activity.

There followed in rapid succession the formation of a Section in Ottawa, Canada, chartered September 5, 1944, and a Section in London, Ontario, chartered on November 24, 1944.

During this period the Canadian Government had passed certain labor laws dealing with collective bargaining, which affected professional engineers and those members of the Institute resident in Canada, who held a professional status and/or held membership in one or more of the Provincial associations of professional engineers, which had been established by Provincial law—Quebec, 1909; Ontario, 1922.

In order to meet this situation, a committee representative of the four major technical societies or bodies in Canada, set up a Committee of Fourteen to meet with the Government to discuss the professional phases of the Federal labor laws. However, prior to this, a representative of the Montreal Section had proposed that the Toronto and Montreal Sections should set up a Council to deal with such problems. Such councils or regional representation having been proposed by ballot to the

membership during 1941, but unfortunately defeated.

In view of the fact that the situation in Canada was unique and because of the fact that the writer was the vice-president of the Institute, and a member of its Board of Directors at the time when these problems became apparent, a request was made to the Board to permit the writer to act for the members of the Institute during this interim period, in collaboration with other engineering societies, in protecting the rights of our members in Canada, in connection with government acts and regulations.

At the same time, recognizing the fact that the members of the Institute resident in Canada were subject to the laws of the Government of Canada and various Provincial acts, the writer placed a motion before the Board of Directors to permit the functioning of national councils on an interim plan of action, which would not involve the Institute financially or legally, to deal with national matters within any one country until such a time as the Board of Directors decide on a final policy, and suggested that the chairmen of the Sections involved should act to give effect to this motion.

The I.R.E. Board of Directors unanimously passed this motion September 6, 1944, but it was not until April 7, 1945, at London, Ontario, that the chairmen met formally to set up a Canadian Council of The Institute of Radio Engineers.

At this meeting, it was agreed that the Council should consist of the chairman of each Section, the immediate past chairman of each Section (as ex-officio members), and to ensure close contact with Headquarters in New York, any Canadian Director of the I.R.E. who is resident in Canada. In the event that no Canadian, resident in Canada, be elected or appointed to the Board, it is further moved that a quorum of the Council be empowered to nominate for membership on the Council a representative for the Canadian membership at large.

It was agreed that the Secretary of the Council should be the chairman or the immediate past chairman of the Section from which the elected chairman of the Council had come, thus ensuring an effective working relationship between these two officers of the Council. The objects of the Council are those set forth in the constitution of The Institute of Radio Engineers, Article 1, Section 2; namely,

(a) Its object shall be the advancement of the theory and practice of radio and allied branches of engineering and of the related arts and sciences, their application to human needs, and the maintenance of a high professional standing among its members.

(b) The Council shall act for the Canadian membership of the Institute on Canadian matters which are the common concern of the Canadian members, but it shall not involve The Institute of Radio Engineers either financially or legally.

(c) The Council shall endeavour to correlate the thoughts and actions of the Canadian membership of the Institute so that it may express the opinions of all Canadian members.

(d) The Council shall consider and take appropriate action in the following: (1) legislative matters affecting the Canadian membership and which are peculiar to Canada; (2) matters which require joint action with other engineering and professional associations in Canada; (3) matters associated with the education of professional workers in the fields of radio, communications and electronics; (4) matters associated with the professional status of workers in these fields; (5) matters associated with the public recognition of the contribution made by the radio and electronic industries in Canada during the war years; (6) such other matters as shall be added by a majority vote of the Council.

(e) The Council shall, at the request of the Sections, serve as a clearinghouse for speakers and papers and in a like capacity in other matters which may arise from time to time.

At the first meeting Dr. F. S. Howes, of the Montreal Section, was named chairman, while E. O. Swan was named vice-chairman. Since this date, L. A. W. East, the chairman of the Montreal Section, was named the secretary, and various committees were established under instructions from Chairman Howes, covering professional status, education committee, and a Canadian charter committee.

The Canadian Council of The Institute of Radio Engineers is a contributing sponsor to the work of the Canadian Radio Technical Planning Board, and maintains membership in the Canadian Council of Professional Engineers and Scientists.

At the present time, the expenses of the Council are defrayed by special vote of the Board of Directors.

It is felt that the work of the Canadian Council will establish a pattern for other groups within the Institute residing outside of the United States, to perform similar functions for their members, and thus further the usefulness and standing of a truly international Institute of Radio Engineers.

A Plea for the Scientific Method*

LOUIS HOFFER†

Summary—**The rapid growth and development of the radio industry accelerated by the impetus of war, the new technical advancements, and the increased complexity of electronic application have indicated the need for more intensive training and a revision of the educational approach in this field.**

Too much emphasis cannot be given to the inculcation of the "scientific habit of thought" and the development of the "scientific research" method in engineering thinking.

What is Scientific Research

SCIENTIFIC research is a method of finding out what we want to know, a technique which has survived because of its utility. The test of the validity of any method lies in its results. Whatever technique achieves that result is necessarily a valid technique. These techniques may vary from the "trial-and-error" method of the simple technician to the rigorous solution of a highly complicated abstract mathematical problem by a professional. The method to be adopted should be the one which is direct, simple, accurate, and reliable.

It is understood that all methods are tentative. Everyone works out something for his own use. Even in the highly formalized sciences, new methods supplant old. Methods in vogue in any field are the residues of techniques used by men to find out what they wanted to know.

However, one method man has used has enabled him to attain an understanding and control of the physical world far beyond the fondest dream of the ancients. In view of the success of this method, the scientific technique, we should be interested in its application to radio engineering. Conscious effort should be made to train the engineer in the common characteristics of this technique.

Critical Discrimination

The first requisite of this method of scientific analysis is the determination and ability to get at the basic facts and not be influenced by mere appearance, prevalent notions, and one's own wishes. Such a mental attitude is the core of the scientific frame of mind. It is the "be all" of all science.

Perhaps the best account of a scientific frame of mind is to be found in Francis Bacon's flattering description of *himself* in one of his famous works: "A mind nimble and versatile enough to catch the similarity between facts, and at the same time, steady enough to fix and distinguish their subtle differences, . . . endowed by nature with the desire to seek, the patience to doubt, fondness to meditate, slowness to assert, readiness to consider, carefulness to dispose and set in order, neither affecting what is new nor admiring what is old . . . "

Critical discrimination is indispensable in all fields and is really a prime requisite in the field of engineering research. The radio industry would profit immeasureably from the scientific frame of mind on the part of every investigation in the laboratory.

Classification

A second requirement of the scientific method is proper classification. Once the data have been collected by observation and experimentation, it is necessary to recognize the relationships that exist between these collected facts. Electrical phenomena should be classified on the basis of the similarity of the character of things, processes, and results achieved.

These classifications may be natural ones, such as the classifications of chemical elements based on atomic weights; or special-purpose classifications for specific objectives. The point is that, in order to arrive at proper conclusions, data must be arranged in some orderly array which naturally leads to such conclusions.

Of course, there is always the danger of improper classification for special purpose. The story is told of that lady who went on a railroad journey with a menagerie of pets. The conductor informed her of the fare for her dogs but didn't know the charge for her other pets. After inquiry of the railroad office, the conductor told the lady that according to the office, "as cats are dogs and rabbits are dogs and so are parrots, the fare would be the same but this here tortoise is an insect so there isn't any charge for it." So far as the railroad was concerned dogs, cats, and parrots belong to the same class, but it would be dangerous to use that type of special-purpose classification in the field of engineering research.

Fair Sample and Verification

Our scientific method starting from observed fact and experimentation usually ends in generalizations of some kind concerning large classes and groups of facts. The right we have to apply the results of our observations of a very limited number of facts to a large number of others which have not been observed depends on the principle of fair sample and verification, which means

* Decimal classification: R070. Original manuscript received by the Institute, October 9, 1945.

† Emerson Radio and Phonograph Corporation, 111 Eighth Avenue, New York 11, N. Y.

that, with reasonable care, it is possible to judge the character of a large group by the aid of a sample selection from it.

Science, in this, as in other respects, follows the lead of the practical man. In business, it is often impossible to examine each item in a large shipment of goods. Hence, businessmen are content to estimate the character of the whole shipment by the aid of a sample. On the whole, experience shows that, provided the sample is selected with some care, it fairly represents the whole, but is itself only a small fraction of the whole. The actual number of samples is less important than the range of selection, and the accuracy of the generalization depends not only on the proper selection but on the degree of accuracy achieved in verifying the conclusion.

Importance of Scientific Technique

Too often only lip service has been paid to the "scientific habit of thought" summarized in the preceding paragraphs. Little effort is made to define the problem and determine the most fruitful line of attack. The game of prodigal waste called "tinkeritis" seems to have a tremendous appeal as an occupational technique, even though the dangers of illusion, fallacy, and "wish fulfillment" are common and serious enough to warn any sensible person who uses the critical approach against excessive confidence in his own conclusions without verification. As a matter of fact, there appears to be a large number of workers in the radio-engineering field who take pride in their ignorance of scientific techniques of logic, and of the value of the faculty of critical discrimination. To them, scientific technique and logical analysis are a kind of refined intellectual game played according to certain rules in selected academic circles.

In reality, however, the scientific method is as important and serious as life itself. Pursued in the right spirit, it will not only lead to the enrichment of engineering research, but will also be an invaluable training in the art of self-criticism which is so necessary to the peace and welfare of humanity.

Need for Pure Research

Scientific research and the scientific frame of mind are handmaidens of pure science. Not only is there a need for the scientific technique in radio engineering, but also a need for fundamental inquiry into basic principles. If only the practical and useful aspects of radio engineering are considered, the radio industry will soon find itself "intellectually handcuffed" and in exactly the same position as that of the manufacturer who suddenly finds himself cut off from his supply of raw materials from which he fabricates his finished product.

As one writer has said, "Everyone recognizes that applied science rests on pure science—that, to cite a single illustration, radio communication would have remained not only impossible but inconceivable save for the fundamental experiments of Faraday, the mathematical formulation of wave theory by Maxwell, the experimental realization of Maxwell's predictions by Hertz—all being advances in knowledge made without thought of practical application or financial return."

"All the ingenuity in the world could not have produced an automobile until science had brought to light the electrical, metallurgical, and mechanical principles involved in its construction."

However, there appears to be a lack of appreciation of the importance and the possibilities of pure research. If insufficient attention and inadequate financing of pure research continues, the stream of creative thought in the field of radio engineering will eventually dry up.

If the Radiation Laboratory at The Massachusetts Institute of Technology (Office of Scientific Research and Development) had such admirable success in war, intelligent industry should see to it that its research role is continued in peace. Industry could certainly develop "a moral-equivalent-of-war spirit" to subsidize a permanent "radiation laboratory" or other research body, devoted to pure science and beneficial to both industry and society.

As Admiral S. C. Hooper said in the June, 1943, issue of the Proceedings of the I.R.E., "The progress and prosperity of our Nation, as well as the readiness of its forces of war, are in no small measure dependent upon keeping peacetime research going full speed."

Radio-Frequency Dehydration of Penicillin Solution*

GEORGE H. BROWN†, FELLOW, I.R.E., R. A. BIERWIRTH‡, MEMBER, I.R.E., AND
CYRIL N. HOYLER†, SENIOR MEMBER, I.R.E.

Summary—**Many pharmaceutical materials, such as penicillin and antitoxins, lose their desirable properties when the material is in liquid solution for a period of time. In general, these solutions suffer from the application of heat. In the dry state, the same materials retain their properties over long periods of time, even at elevated temperatures. Drying of penicillin is generally carried out by freezing and then applying a high vacuum so that the material is dried by sublimation. This paper describes certain experiments and developments made by the authors to bring about the drying of penicillin by the application of radio-frequency power in a very moderate vacuum.**

The system presented in this paper has been divided into three parts: (1) bulk concentrating of the solution; (2) drying the concentrate in the final containers to a moisture content of a few per cent; and (3) completing the drying in auxiliary chambers. The equipment is capable of producing 2000 dry bottles per hour, using inexpensive vacuum pumps and simple condensers. The whole system operates with a minimum of maintenance.

A few experiments have been made concerning the drying of other pharmaceuticals with promising results.

I. Introduction

LATE IN 1943, many newspaper and magazine articles were published which told the story of the manufacture of penicillin. Particular emphasis was given to the problem of drying the product for packaging. Two points were stressed. First, penicillin in a water solution soon lost its bactericidal properties so it was necessary to dry the product thoroughly before storage or shipment. Second, penicillin solution was so sensitive to heat that it was necessary to keep the solution in a frozen condition before drying and even to dry in the frozen state.This freeze-drying process was expensive and took a long time. Our previous work in dehydration of materials by means of radio-frequency power, particularly the drying of rayon cakes in a vacuum, led us to speculate about the possibilities of contributing something to the penicillin production program by the application of radio-frequency power. The writers then suggested to the late Dr. George A. Harrop, Director of Squibb Biological Laboratories, that it might be possible to evaporate the water from the liquid state if the vacuum were such that the boiling point remained at about 50 degrees Fahrenheit and the energy necessary for vaporization were supplied by a radio-frequency generator. Dr. Harrop expressed great interest in the possibilities and offered to co-operate in tests to determine whether such a procedure would harm the product. Before such tests could be made, other producers of penicillin heard of our ideas and wished to discuss the proposal. Some of these producers provided samples for test purposes. A description of the pertinent tests will be included in this paper. In order to orient the reader, it seems desirable to say a few words concerning the methods of testing for bactericidal properties and the means of expressing these properties, since the purpose of our early experiments was to determine how our radio-frequency methods affected these properties.

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II. Penicillin Activity or Potency

Pure crystals of penicillin salt are believed to be very stable and to have constant bacteria-destroying properties for a given weight. Pure crystals are difficult to obtain. However, in limited amounts, these crystals are useful in testing solutions of penicillin which contain impurities and are of unknown activity. The Oxford unit of activity was selected shortly after the discovery of penicillin, so the unit is smaller than necessary for convenient use.[1] Pure crystals of penicillin possess 1667 Oxford units of activity per milligram.

In testing a solution of unknown potency, various dilutions of the substance are placed on a bacteria culture. At other spots on the same plate, a penicillin solution of known activity is placed. After incubation for a fixed period of time, observations are made of the dimensions of the zones of inhibition.

Another test makes use of observations of turbidity of various dilutions of penicillin in cultures of bacteria. Again the observations are correlated with a known solution.

The accuracy or consistency of these bacteriological determinations is not very great. A number of measurements of potency taken on the same sample may show deviations as high as plus or minus fifteen per cent from the average.

In our experiments, control samples were kept so that any deterioration of the solution with time could be sorted out from the effect of radio frequency.

III. Initial Experiments with Radio-Frequency Heating

With a few exceptions, the people connected with penicillin production assured us that it was extremely important to keep the solution frozen throughout the drying cycle. They warned that if this were not done the potency would be lost. The statements did not appear to be based on experience related to the method of

[1] One Oxford unit is that amount of penicillin which, when dissolved in 50 milliliters of meat-extract broth, just inhibits completely the growth of the test strain of *Staphylococcus aureus.*

drying which we proposed. Therefore, it seemed desirable to determine by experiment just what the effect would be.

To dry by conventional freeze-drying methods,[2] it is necessary to maintain a vacuum of between 100 and 300 microns (0.1 to 0.3 millimeters of mercury). We were not interested in freeze-drying with radio-frequency power, for a number of reasons. In the first place, if we were to contribute to the penicillin program by furnishing a faster and simpler system, the elimination of the expensive high-vacuum systems, complicated condensers, and elaborate refrigeration seemed to be the most important step. Second, for the same voltage or electric intensity, the residual atmosphere ionizes much more readily with a vacuum of 100 microns than it does at a vacuum of about 20 millimeters, which is the order of vacuum that we had in mind for our purpose. In addition to the fact that the air ionizes easier at 100 microns, the solution is frozen when under this vacuum. The electrical conductivity of frozen penicillin is so much lower than that of the liquid that much higher voltages must be used to generate the same power. This effect is very pronounced. When attempts were made to dry in the frozen state with radio-frequency power, it was necessary to reduce the radio-frequency voltages to such an extent that it became apparent that most of the heat of vaporization was being supplied by heat conduction from the warm air of the laboratory through the glass walls of the container.

A number of samples of a weak solution of penicillin were submitted by the representatives of a penicillin producer. They were anxious that the tests be made in the frozen state, so only one sample was subjected to the test which we wished to make. Each sample was a frozen solution weighing 1000 milligrams. The results of the experiments are given in Table I.

TABLE I

Sample	Final Weight (Milligrams)	Units of Activity Per Milligram	Total Units In Sample	Treatment
Control	58.7	67.0	3933.0	Control dried by conventional freeze-drying.
1	59.1	47.0	2780.0	Shelled[1] and dried for 80 minutes with radio frequency, vacuum of 80 microns.
2	68.4	55.0	3760.0	Shelled[1] and dried for 60 minutes with radio frequency, vacuum of 100 microns.
3	57.5	76.0	4370.0	Not shelled. Radio frequency applied for 10 minutes, but vacuum of 100 microns for 90 minutes.
4	55.5	82.0	4550.0	Frozen when placed on pump, and vacuum of 10 millimeters applied. Radio-frequency power used. Melted immediately. Appeared dry in five minutes. Heating continued for total of 30 minutes.

[1] Shelling is the procedure of freezing the liquid in a thin layer on the sides of the bottle, accomplished by holding the bottle on its side and slowly rotating the bottle while submerged in a refrigerant until the contents are completely frozen.

The tests shown in Table I indicated that the best results were achieved by boiling out the water at the relatively high pressure of 10 millimeters. Since this was not in accordance with the opinion of experts, it seemed desirable to verify this conclusion by further tests. Accordingly, four samples obtained from E. R. Squibb and Sons were treated. These samples were all frozen when placed in the drying apparatus. However, the pressures were high enough so that the material quickly melted. The following conditions prevailed for these samples:

Sample 1: Vacuum of 10 millimeters. Drying time 52 minutes.

Sample 2: Vacuum of 18 millimeters. Drying time 103 minutes.

Sample 3: Vacuum of 6 millimeters. Drying time 75 minutes.

Sample 4: Vacuum of 3 millimeters. Drying time 35 minutes.

The activity tests reported by Squibb Laboratories were as shown in Tables II and III.

TABLE II

	Control No. 1	Control No. 2
First dilution	73,100 units	82,400 units
Second dilution	70,900 units	82,500 units
Average	72,000 units	82,450 units
Average of controls	77,225 units	

TABLE III

Sample	First Dilution	Second Dilution	Average
1	82,400	76,200	79,300
2	83,560	82,100	82,830
3	76,100	77,500	76,800
4	75,000	72,700	73,850
Average of radio-frequency samples			78,195

Four more samples were later dried under the following conditions and tested for activity by Squibb Laboratories (Table IV).

TABLE IV

Sample	Activity	Treatment
5	88,644	Vacuum of 150 microns, with radio frequency. Sample frozen throughout treatment. Drying time 5 hours.
6	87,316	Vacuum of 10 millimeters. Drying time 70 minutes. No loss of weight after 40 minutes. Sample melted quickly and had dried appearance after six minutes.
7	83,664	Vacuum of 10 millimeters. Drying time 60 minutes. No loss of weight after 45 minutes. Sample melted quickly and had dried appearance after six minutes.
8	75,862	Vacuum of 30 millimeters. Drying time 60 minutes.
Control	70,625	

The people at the Squibb Laboratories attributed the lower activity of sample 8 and the control to the fact that both specimens had been kept in solution in a frozen state for several days after samples 5, 6, and 7 were dried.

In any event, we felt that the test results showed that no harmful effects appeared when the dehydration took place with vacuums of the order of 10 millimeters. These initial tests have been repeated many times, always with the same result—no loss in potency.

In the above tests, it may appear that it is possible to freeze-dry with high vacuum in times varying from one

[2] Earl W. Flosdorf, Lewis W. Hull, and Stuart Mudd, "Drying by sublimation," *Jour. Immunology*, vol. 50, pp. 21–54; January, 1945.

to five hours. In our laboratory tests, the sample was in a thin-walled glass bottle which was connected to a manifold. Here the entire bottle was exposed to the warm air of the laboratory. In practice, where thousands of these bottles are placed in large vacuum tanks, no circulation of warm air takes place. The drying time by this method may take from eight to twenty hours.[3]

IV. A Radio-Frequency Bulk Reducer for Pencillin

As a result of these initial successes, much interest was exhibited by a number of pencillin producers. It seemed extremely desirable to review the work and consider how best to fit radio-frequency power into the picture.

After the penicillin mold is grown and the penicillin harvested, the material goes through a number of complex chemical processes from which it finally emerges as a weak water solution of a sodium or calcium salt of penicillin, with varying amounts of impurities. At the time that we were interested in considering the situation, it was the practice to freeze or shell this solution on the walls of a number of large glass bottles which were then attached to a manifold of a vacuum system. The vacuum was maintained at about 100 microns. The only heat supplied was by conduction through the walls of the bottles. After many hours, the material was considered to be dry and was then scraped from the bottles and pulverized. A measurement of potency, in Oxford units per milligram, was then made to determine the amount by weight that would be placed in the final ampoule or bottle. Since it was necessary to maintain a high degree of sterility, this operation was done in an air-conditioned box containing an analytical balance, a tray of penicillin powder, a rack of bottles, and the necessary loading tools. The operator manipulated the apparatus by inserting his arms into a pair of rubber gloves that were anchored to two openings in the box and by peeping through a small glass window in the wall of the box. This process was slow and expensive, as well as tedious and inaccurate.

We were told that most processors were considering a modification in which freeze-drying in the large bottles would be halted at a point where the activity of residue would be between 10,000 and 100,000 units per cubic centimeter. The material would then be allowed to melt and the liquid would be assayed for potency. Depending on the activity attained, between one and ten cubic centimeters of material could be accurately measured into each final container. Then a rack of final containers would be frozen and inserted in a large vacuum chamber and the pressure reduced to about 100 microns, for the final drying. Because of this change of procedure, it was decided that a radio-frequency bulk reducer was first in order of importance. The bulk reducer could then be used to remove sufficient water to concentrate to a potency of 100,000 units per cubic centimeter, after which one cubic centimeter of the material would be measured into each of the final containers and taken down to complete dryness by the freeze-drying high-vacuum chambers. The twofold advantage of the radio-frequency bulk reducer is that the expensive high-vacuum bulk reducer and refrigeration system is eliminated and the output of the final freeze-drying cabinets is increased because only one cubic centimeter of water is removed from each final container.

The first experiments in bulk reduction were conducted with a laboratory oscillator which could be changed readily to a number of frequencies. The frequencies investigated particularly corresponded to the frequencies of commercially available oscillators.

The experimental arrangement is shown in Fig. 1. The penicillin was contained in the large bottle shown at the table's edge. The tall chimney was necessary to prevent losses of material due to splashing. The electrodes were simply thin sheets of metal taped to the outside surface of the bottle. The bottom of the bottle was made concave so the load presented to the oscillator would remain fairly constant throughout the concentration cycle. This enabled us to use wider electrodes and still preserve this constancy of load. With a flat bottom, the liquid was sometimes concentrated below the level of the electrodes so that there was danger of ionization in the air space and sometimes arcing from the electrodes.

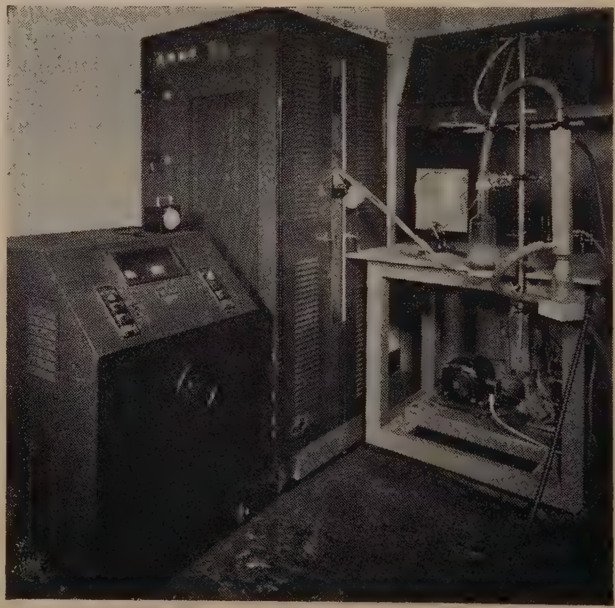

Fig. 1—First experimental bulk reducer.

Attempts were made to operate at a frequency of 10 megacycles, but we were not able to overcome sparking at the electrodes. Further experiment showed that, at a frequency of 28 megacycles, there was little danger of electrode trouble, provided the concave bottom was used.

[3] See p. 38 of footnote reference 2.

A Beach-Russ vacuum pump, of the wet type, with a capacity of six cubic feet per minute, was used through-

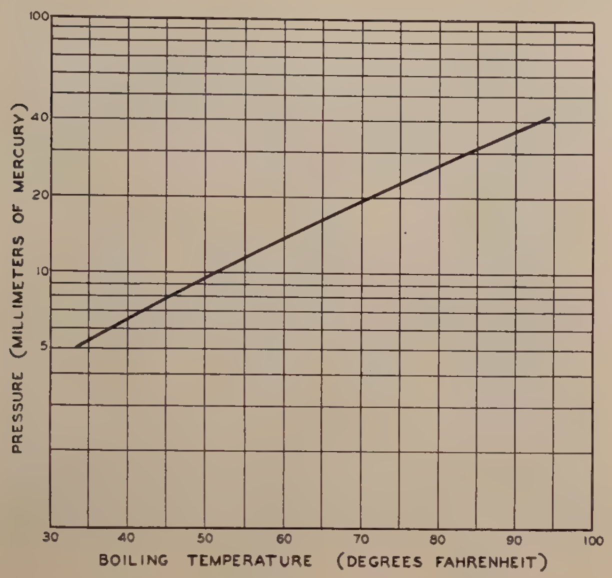

Fig. 2—Boiling temperature versus pressure.

out the experiments with the bulk reducer and in the final model. The water vapor was passed into a water-

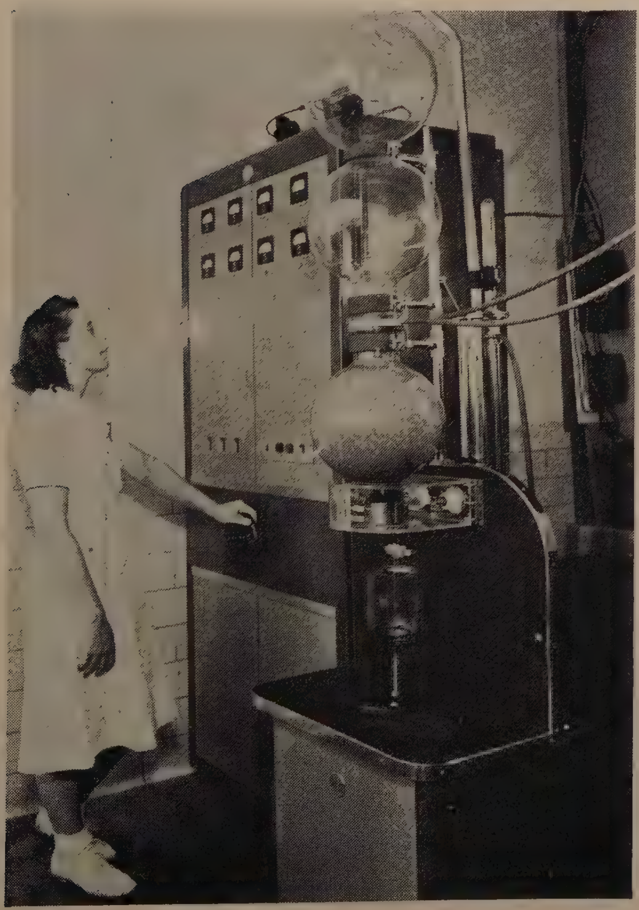

Fig. 3—Pilot bulk reducer in operation at Squibb Laboratory.

cooled condenser so the pump was not forced to handle large quantities of vapor.

A large number of experiments was carried out to learn the necessary technique for a practical system. Further tests were necessary to determine the safe range of temperatures and pressures which could be used without harming the penicillin. The temperature at which water boils as a function of pressure is shown in Fig. 2. While much of our operation took place at a pressure

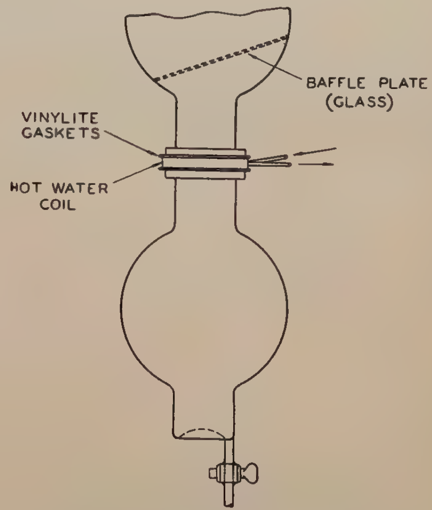

Fig. 4—Diagram of bottles with baffle plate and water coil in place.

between 10 and 20 millimeters of mercury, we have run the equipment at pressures up to 40 millimeters without loss in potency of the penicillin.

A pilot unit was next constructed. The oscillator consisted of two RCA-833A vacuum tubes acting as a conventional oscillator at 28 megacycles, with a power output of two kilowatts. Fig. 3 shows the equipment in operation at Squibb's plant. The three large glass bottles provide room for foaming of the liquid when the vacuum is applied. Penicillin solution is extremely foamy so that our efforts in suppressing the foam were extensive. A baffle plate was inserted in the middle bottle. Also, a coil made of silver-plated copper tubing was inserted between the bottom and the middle bottles. The arrangement is shown in Fig. 4. Warm water (125 degrees Fahrenheit) was passed through this coil. When the foam and bubbles rise and contact the warm coil, the part of the bubble in contact with the coil evaporates and the bubbles collapse. This arrangement has been extremely successful in the operation at the Squibb laboratory.

In connection with the foaming problem, it might be mentioned that the excessive foaming difficulties were encountered at a time when most penicillin solution carried a large amount of impurities. Observations made recently show that the penicillin solution now being produced by most manufacturers has improved in characteristics to such an extent that neither the baffle plate

nor the hot-water coil need be used. Actually, the most violent foaming only fills the lower half of the bottom flask.

In using the bulk reducer, the sequence of operations is as follows: The hot-water supply for the bubble-breaking coil and the cold-water supply for the condenser are turned on and the vacuum pump started. When a liter of material is ready for treatment, a flask containing the solution is placed beneath the bottom bottle and the pet cock is opened. The penicillin solution is thus sucked up into the bottle. The pet cock is then closed and the oscillator turned on. With full power applied, 1000 cubic centimeters of liquid will be reduced to 100 cubic centimeters in eighteen minutes.

After turning off the oscillator, the vacuum pump is stopped and the relief valve is opened. Then the pet cock below the bottom bottle is opened and the penicillin solution drained from the bottle. After this operation, the pump is started and a new supply of penicillin is drawn into the bottle.

V. Drying in the Final Container

In Section III, the results of early experiments were reported. Here the drying was carried out in large test

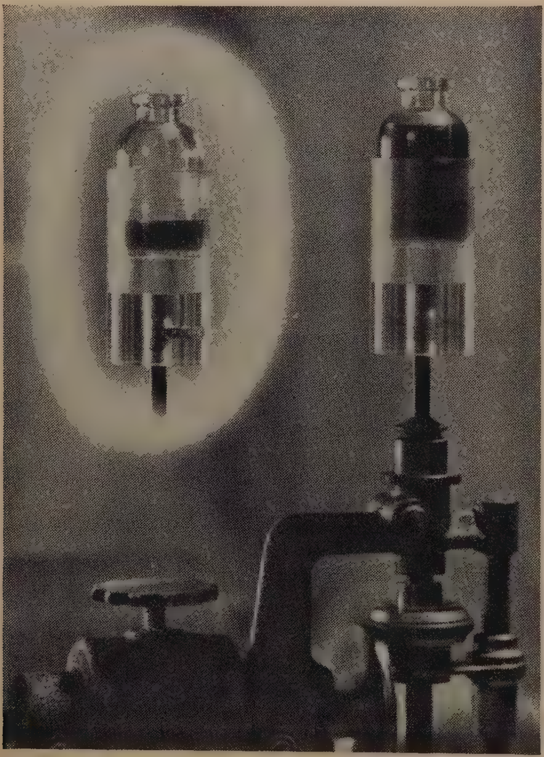

Fig. 5—The bottle is placed in a plastic cup. The bottle at the right is rotating at a speed of 3000 revolutions per minute. The liquid forms a thin film on the vall of the bottle.

tubes or ampoules. The material spattered up on the walls of the container when the vacuum was applied and spattered again when the radio-frequency power was turned on. This resulted in a messy appearance when finally dried. However, we were interested in the effect on the activity and for the moment were not concerned with appearance. When later attempts were made to

Fig. 6—Samples of penicillin dried by spinning in a radio-frequency field. Each bottle contains 100,000 units of penicillin, with varying amounts of impurities.

dry in the final bottle, it became apparent that it was almost impossible to keep all the material in the bottle. Initial application of the vacuum usually resulted in loss of a certain amount of solution due to spattering. After the application of power, severe bumping sometimes entirely emptied the bottle. Numerous expedients were resorted to in an attempt to solve this problem. Rotating the bottle at high speed during the evacuating and drying period revealed that this was a way to eliminate these difficulties. Fig. 5 illustrates this action. The bottle on the left contains about one cubic centimeter of penicillin solution. The same bottle is shown on the right, rotating at a rate of 3000 revolutions per minute. The liquid and solids in the liquid are then formed in a thin layer on the walls of the bottle. The force on the solids at this rotational speed is estimated to be at least one hundred times the force of gravity. When vacuum is applied, no material is lost. Then, as radio-frequency power is applied, the material dries in a thin film on the side of the bottle with a rather pleasing appearance. Dried samples may be seen in Fig. 6. A single-unit drier consisting of a plastic cup driven at a speed of 3000 revolutions per minute and a set of electrodes is shown in Fig. 7. This unit was found to be very useful in the laboratory, and several hundred samples of penicillin solution have been dried under various conditions with this equipment. With approximately 20 watts of power at 30 megacycles, it is possible to remove most of the moisture from one cubic centimeter of solution in three minutes. Typical data taken during a drying cycle is shown in Fig. 8. It may be seen that most of the water is removed in the first 60 seconds. The rate of drying

progressively decreases as time passes. Measurements of the Q factor of the circuit taken during the drying

Fig. 7—A laboratory arrangement for drying one bottle at a time.

cycle are also shown in Fig. 8. As the material dries out, the Q becomes high so that there is quite a change in loading of the oscillator during the drying period. At the

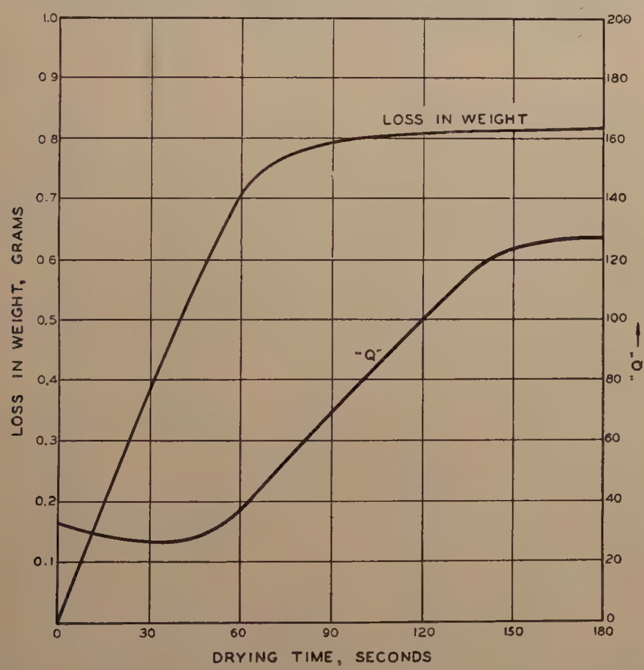

Fig. 8—Drying rate and Q of a single bottle as a function of time.

end of three minutes, the moisture content is reduced to about 4 per cent. Application of radio-frequency power for longer periods will reduce the moisture content still further. However, by the time the moisture content has been reduced to 4 per cent, the material is very stable and will stand much higher temperatures than it does when in the liquid condition. This allows us to introduce another step which is important in the construction of final equipment, since it permits the use of a minimum number of rotating cups. The bottles are transferred to metal plates which are heated by ordinary electric heaters. A bell jar is then placed over the heater plate and the system evacuated. Since the ma-

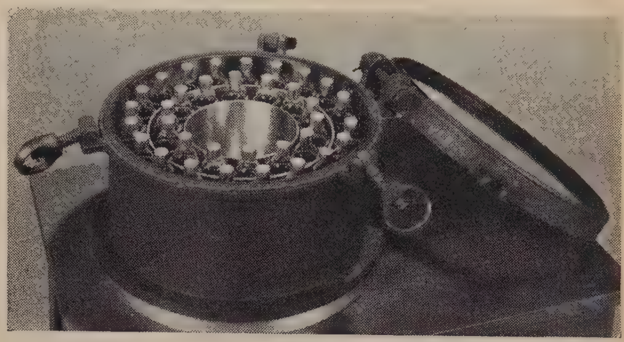

Fig. 9—An experimental drying chamber which holds thirty-four bottles.

terial is dried on the sides of the bottle in a thin film, heat conduction through the glass bottle to remove the last traces of moisture is efficient.

Early in the investigation, several samples of penicillin solution were dried and tested for activity and moisture content. In each case the bottle was rotated for three minutes in a radio-frequency field, with a vacuum of 50 millimeters of mercury. Then the bottles were removed from the radio-frequency field and placed in a metal box which in turn was placed in a bell jar, with a vacuum of 8 millimeters of mercury. An electric heater, thermostatically controlled, heated the metal plate which supported the bell jar to a temperature of 65 degrees centigrade. At the end of the run, the bottles reached a temperature of 50 to 55 degrees centigrade. The conditions of test as well as activity measurements are show in Table V.

TABLE V

Sample	Time in Bell Jar	Activity Oxford Units	Number of Activity Tests
Control	—	95,954	8
Control	—	103,441	6
204	2 hours	115,233	4
205	2 hours	101,853	4
206	2 hours	101,561	4
212	1 hour	108,265	4
214	1 hour	110,349	4

The moisture determinations were as shown in Table VI.

TABLE VI

Time in Bell Jar	Per Cent Moisture (Average of Three Samples)
2 hours	0.43
1 hour	0.60
Zero	4.13

With these and similar data at hand, the construction of a system which would handle large numbers of bottles was started. A unit of the system is shown in Fig. 9. A port-hole frame is used as a vacuum chamber. Thirty-four plastic cups are mounted on the peripheries of two

circles. An electrode, insulated from ground may be seen between the two rings of bottles. An inner metal ring and the wall of the port form the two ground electrodes. Radio-frequency power is fed in through a vacuum-tight insulating bushing. Each of the plastic cups is individually gear driven at a speed of 3000 revolutions per minute.

Fig. 10 shows a circular table which supports six of these chambers. Each chamber is subjected to radio-frequency power for three minutes. A rotary switch is used so that three chambers are connected to the oscillator at one time. One chamber is in the wet stage; that is, it is in the first third of its drying cycle. Another chamber is in the second third of its drying cycle, while a third chamber is in the last stage of the drying cycle. At the end of one minute, the chamber containing dry bottles is switched from the oscillator and a fresh chamber is connected in the group. Thus the oscillator remains essentially fully loaded at all times. An examination of Fig. 8 reveals the reasoning behind this procedure. In addition, the automatic load-tuning circuit incorporated in the RCA-2B oscillator insures complete stability with respect to power output. While three chambers are being dried, the fourth is in the process of unloading, the fifth is being loaded, and the sixth is in the initial stages of evacuation. Since one chamber of thirty-four bottles is unloaded every minute, over two thousand bottles are processed each hour.

Fig. 10—A circular table with six drying chambers. This equipment dries 2000 bottles per hour. The infrared lamps prevent condensation on the cold lids of the vacuum chambers.

The bottles are next loaded into aluminum trays and the trays are placed on the electrically heated bases shown in Fig. 11. The metal domes are then closed and evacuated to about 5 millimeters of mercury. After one-half hour in these auxiliary domes, the bottles are removed for capping.

The equipment shown in Fig. 10 is largely automatic. The operator's duties consist of operating the vacuum applicator valves and vacuum release valves on the side of the table, as well as opening the lids of the chambers, removing the processed bottles, and inserting new bottles to be dried. When the operator closes the lid to a chamber, a microswitch is actuated and the bottles begin rotating. A green pilot light at each chamber indicates that the chamber may be opened. A red pilot light shows that radio-frequency power is being applied to the chamber. At the end of each minute, a timer turns off the oscillator, operates the rotary switch so that one chamber is dropped out and a fresh chamber connected, and then turns on the oscillator. If the operator has not yet processed this fresh chamber and thrown the vacuum-applicator valve to the final position, the oscillator will not turn on, and an amber pilot light will indicate the chamber that is delaying the proceedings. When the vacuum-applicator valve is finally thrown, the one-minute cycles will then be resumed.

Fig. 11—The auxiliary drying chambers.

VI. Electrical Conductivity of Penicillin

It has been suggested many times that we reconsider the situation and attempt to dry in the frozen state. Since we have shown that our method of drying produces a dry product with no loss in activity, there seems to be no point to the introduction of greater complications of higher vacuums and colder condensers. In addition, the high voltage required in the presence of the more perfect vacuum seems to be an obstacle which will not be solved easily. Early in this paper, it was stated that the voltages necessary to develop power in the frozen material were markedly higher than for the liquid solution. This point may be illustrated by a specific example.

The power generated in a one-centimeter cube of material is

$$P(\text{watts}) = E^2\sigma$$

where σ = conductivity of the material in mhos for a centimeter cube

E = electric intensity in the cube in volts per centimeter.

Then for one watt generated in the cube, the electric intensity is

$$E = \frac{1}{\sqrt{\sigma}}.$$

This electric intensity may be regarded as the voltage required between opposing faces of the cube.

Murphy and Morgan[4] show that the conductivity of ice increases as a function of frequency but flattens off completely when the frequency is greater than 10,000 cycles. For ice at -0.8 degrees centigrade, they show a conductivity of 4.15×10^{-7} mhos for a centimeter cube. When substituted into the above equation, it is seen that the voltage across the one-centimeter cube for one watt is 1550 volts. However, this temperature is very close to the melting point of ice at atmospheric pressure. A few degrees drop in temperature reduces the conductivity appreciably. At -45.8 degrees centigrade, a temperature where freeze-drying is usually carried out, the conductivity has dropped to 10^{-8} mhos for a centimeter cube, with a resultant field intensity of 10,000 volts per centimeter.

Measurements on typical penicillin solution in the liquid state (20 degrees centigrade) show that the conductivity may be expressed as

$$\sigma = 10^{-6}A$$

where A is the activity measured in Oxford units per cubic centimeter. Thus when the activity is 100,000 units per cubic centimeter, the conductivity is 0.1 mhos for a centimeter cube, and the field intensity necessary to generate one watt in a centimeter cube reaches the remarkably low value of 3.16 volts per centimeter.

In addition to these quantitative relations, the difficulties of coupling radio-frequency energy to ice may be realized by means of a simple experiment. A set of electrodes which will hold a penicillin bottle is mounted on the capacitor terminals of a conventional Q meter. Then if a bottle containing several cubic centimeters of frozen penicillin solution and an empty bottle are successively plugged into the electrodes, it is impossible to detect a difference in the reading of the Q meter.

[4] E. J. Murphy and S. O. Morgan, "The dielectric properties of insulating materials," *Bell Sys. Tech. Jour.*, vol. 18, p. 512; July, 1939.

VII. Conclusion

Experiments leading to the development of a successful method of dehydrating penicillin solution by means of radio-frequency heat have been presented. The process has been divided into three parts: (1) bulk reducing of the solution to achieve an activity of 100,000 units per cubic centimeter; (2) drying one cubic centimeter of the concentrate in the final bottles until the moisture content is reduced to four per cent; and (3) completing the drying in auxiliary chambers. The equipment described is capable of producing 2000 dry bottles each hour, using inexpensive vacuum pumps and simple condensers. The whole system operates with a minimum of maintenance.

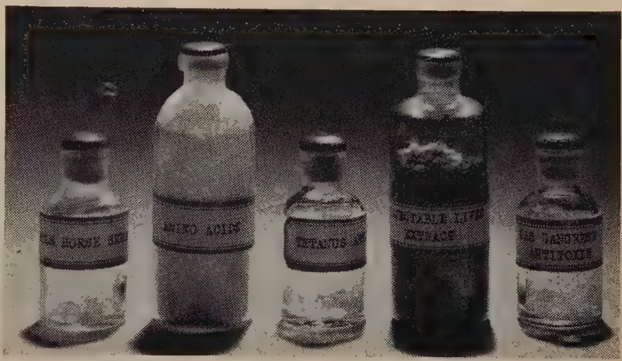

Fig. 12—Representative pharmaceuticals dried by the radio-frequency method. From left to right: whole horse serum, amino acids, tetanus antitoxin, injectable liver extract, gas gangrene antitoxin.

When the bottles are removed from the auxiliary heaters, they are closed with rubber stoppers, and aluminum caps are pressed on to hold the stoppers in place. The bottles may then be stored for long periods of time until needed. At that time, sterile water is injected through the rubber stopper with a hypodermic syringe, the penicillin readily dissolves in the water, and the solution is withdrawn into the syringe.

Because the techniques described in this paper are so different from conventional methods of drying of pharmaceuticals and biologicals, it is difficult to predict what will happen to the many products which might be dried by these methods. Until a broad experience has been built up, it is necessary to submit each product to experiment. The authors have dried a limited number of products with results which show great promise. Representative samples are shown in Fig. 12.

A Vacuum-Contained Push-Pull Triode Transmitter*

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Summary—**A 600-megacycle transmitter is described which varies from the usual triode design in that the resonating grid and plate circuits are contained in vacuum and form an integral part of the grid and plate structures. The design described, while applicable for continuous-wave operation, covers particularly United States Army-type tube VT-158 constructed only for pulse operation. Peak powers of 200 to 300 kilowatts can be obtained under proper conditions of operation.**

I. Introduction

A FEW YEARS ago Major-General Roger B. Colton, then Director of the Signal Corps Laboratories, Fort Monmouth, N. J., noticed how much difficulty was being experienced with radio-frequency sparking and loading of more or less conventional ultra-high-frequency high-power, triode oscillator circuits. He suggested that most of the troubles would be eliminated if the oscillators were constructed with the circuits contained in a vacuum. This paper will describe a transmitter which was subsequently built so as to incorporate most of the radio-frequency circuits inside the vacuum envelope.

II. Design Description

The illustration represents a tube constructed by the authors which can be made to oscillate in a narrow frequency band between 200 and 700 megacycles. Although satisfactory continuous-wave oscillators of this type have been built, the main application has been in the field of high-power pulse oscillators capable of furnishing about two- or three-hundred kilowatts of radio-frequency power.

Fig. 1 shows a general view of one model of this type of tube. The two vertical, parallel rods at the top of the tube form a balanced-line output circuit which is coupled directly to the plate circuit. The parallel transmission line serves in a rough way as a matching circuit to connect the tube to a 50-ohm concentric line, such as is often used to transmit power in this frequency band.

The uppermost loop, or plate loop, is connected directly to the anodes in such a way as simultaneously to reduce the problem of connecting the oscillating circuit to the plates, and by virtue of the flat surface of the loop, increase the effective radiating surface and power dissipation of the plates. The plate loop, made up of two U sections, serves to maintain the symmetry of the circuit, and of course each of the two U sections, in effect, resonates with half of the interelement capacitances. Thus for a given half-loop length and set of interelement capacitances, the tube may be made to oscillate at a higher frequency than if only one half loop was used.

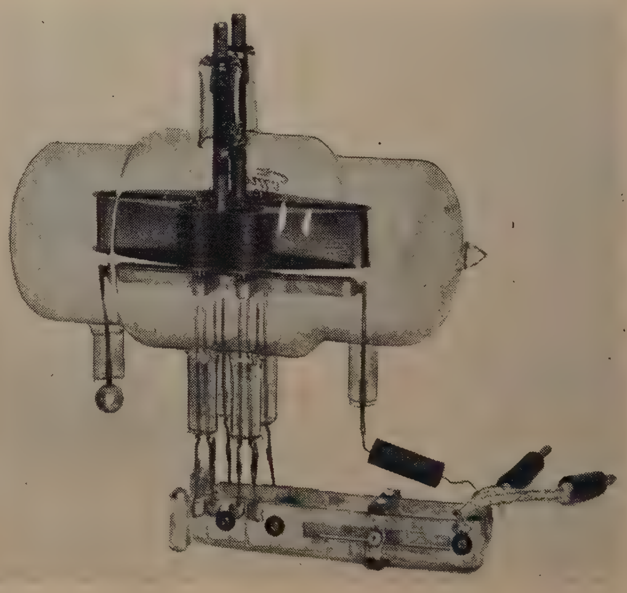

Fig. 1—Front view of transmitting tube.

It has been found advantageous to make both the upper and lower loops of tantalum, since this metal is relatively easy to work and is well known as a good getter. Production models of this general design depend entirely on the tantalum for getter action.

A word should be said about the two anodes in parallel, which in turn operate in push pull with two more anodes in parallel. Several successful continuous-wave oscillators were built with single anodes on each side, but when the tube was pulsed to obtain peak powers several thousand times larger than the average power, it was found that the main limitation in tube output was determined by the amount of available peak emission. The parallel-anode type of construction thus doubled the available peak emission for a given type of element configuration.

The lower loop is the grid loop. Its primary purpose is to obtain grid driving power from the plate circuit. It has been found that relatively little of the power dissipated in the grid cages is conducted along the grid

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† Signal Corps Engineering Laboratories, Bradley Beach, N. J.

wires to the external loop, and therefore it can have less radiating surface than the plate loop. In normal operation, the anodes are run at a red heat, and the filaments are run at a somewhat higher temperature than is usually used. This results in having the grid cages located between two very hot elements, and they are therefore very sensitive to grid contamination and subsequent blocking during oscillation. Under such conditions when plain tantalum wire is used in the cages, the tube becomes sufficiently contaminated to be inoperative in about 24 hours. This difficulty has been overcome completely by the use of Eimac "X" grid cages,[1] and without exception the end of tube life is determined by loss of emission, as it should be, after many hundreds of hours.

high voltage applied to the filaments. The filament line is tuned so that the filaments are effectively points of zero radio-frequency voltage. It has been found that this external line may be either parallel or at right angles to the axis of the tube. While it is usually preferable to have the filament line at right angles, since radio-frequency corona is less troublesome, where space is important it is mounted parallel with the axis of the tube.

Four filaments, instead of two larger ones, were used since several advantages were thus obtained. The filament spirals could be made more compactly and therefore less susceptible to sagging, and the filament power could be dissipated in two anodes instead of one, with attendent lower element temperatures. This construction also allows closer spacing of the elements, an important advantage at these frequencies where transit time must be considered. The obvious disadvantage of this multiple-filament design is its complexity, and it must be said that the highest tribute is due the several tube manufacturers for their ingenuity in overcoming the production difficulties by the development of very clever techniques, tools and jigs, and particularly to Mr. W. Eitel and Mr. J. A. McCullough of Eitel McCullough, Inc., for early assistance in establishing the mechanical design of the production-type tube.

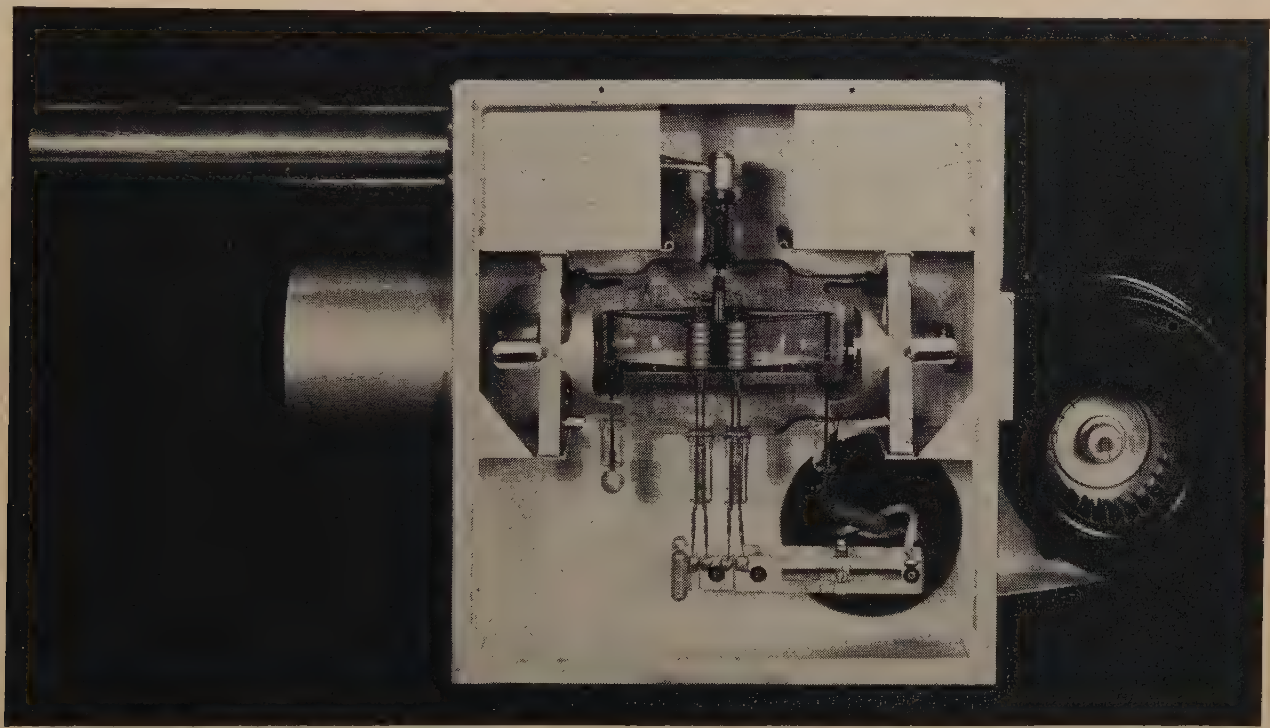

Fig. 2—Front view of transmitting tube and shield.

The filaments are made of carbonized, thoriated tungsten wire and run at a temperature which gives an emission of about 200 milliamperes per watt. In order to obtain satisfactory operation, an external, parallel transmission line is attached so that the two filaments on the same side of the tube are tied together for radio-frequency considerations. Several such transmission lines have been designed with which it is possible to run all four filaments in series if a higher voltage, lower current source is to be used. Since it is most convenient to connect the plate output directly to the output transmission line, the tube is usually run with the plate circuit at direct-current ground potential, and a negative

[1] H. E. Sorg and G. A. Becker, "Grid emission in vacuum tubes," *Electronics*, vol. 7, pp. 104–109; July, 1945.

III. Shielding

If the oscillator is used without a shield, about 80 per cent of the power output obtainable with a shield is

realized. In general, the shielding found most successful is shown in Fig. 2 and consists of a rectangular metal box with two central transverse shelves which serve to hold the tube, and to separate the cavity containing the tube from the cavities containing the output and filament circuits. The output cavity has been designed to act as a "bazooka" or radio-frequency choke at the end of the concentric line, and thus accomplish the transition from the balanced output of the tube to the unbalanced, concentric transmission line. The filament-line cavity serves primarily to contain the radiation, but its shape also determines the magnitude of standing waves at the point where the filament pins are sealed through the glass envelope.

In contrast with the situation at the plate output seals, the filament leads inside the envelope act as a rough transformer to produce a high radio-frequency voltage at the filament seals. At extremely high voltages and powers, this causes corona and subsequent detuning of the oscillator in an erratic manner. This can be avoided for peak powers greater than 300 kilowatts by use of a vertical, external filament line and proper design of the cavity. The frequency of the oscillator does not seem to be especially sensitive to the type of shield used.

IV. Operating Characteristics

A typical set of characteristics is shown in Table I. (Joint Army-Navy terminology is used.)

Table I

Description: Ultra-High-Frequency Pulsed Oscillator VT-158.

Ratings:	E_f	I_f	E_b	E_c	I_b	I_c	P_p	p_i	Modulation
Absolute:	volts	amperes	kilovolts direct current		amperes	—	watts	kilowatts	
Maximum:	10.5	10.5 (Note 1)	30 (Note 2)		70	—	400 (Note 3)	1500	Plate
Test conditions:		10.2	—	—	—	—	—	—	

Reference	*Test: Test Conditions*		*Acceptance Limits* Minimum	*Acceptance Limits* Maximum	
F-6b(3)	*Bump: Angle = 10 degrees (Note 4)				
F-6b(4)	*Bump: Angle = 3½ degrees (Note 4)				
—	Filament voltage (Note 1)	E_f:	9.8	10.2	volts
—	†Emission: $e_b = e_c = 3000$ volts	i_s:	70	—	amperes
—	†Pulse operation $E_b = 20$ kilovolts direct; current pulse repetition rate = 240 pulses per second $F = 595$ to 600 megacycles, pulse width = 1 microsecond	P_o:	150	—	kilowatts
F-6p	*Capacitance:	C_{gp}:	11.8	16.0	micromicrofarads
		C_{gf}:	11.9	16.1	micromicrofarads
		C_{pf}:	1.25	1.69	micromicrofarads
—	*Gas Test: pulse operation: (Note 5)				
F-4	Life Test: pulse operation:		500	—	hours
F-4B	Life-Test end point:	P_o:	135	—	kilowatts

Note 1: Measured for each filament separately.
Note 2: The tube is self-biased (80 ohms 5W, grid resistor).
Note 3: A minimum of 60 cubic feet of free air per minute is required across the tube for a plate dissipation of 400 watts.
Note 4: The hammer arm shall be allowed to strike the glass envelope at an angle of 90 degrees to the plane of grid leads.
Note 5: The tube shall be subjected to the pulse-operation test three times for two minutes at two-minute intervals applying all voltages including filament voltage and cooling devices, simultaneously. The temperature of the bulb of the tube shall not exceed 50 degrees centigrade at start of this test. There shall be no indications of gas or seal failure during or at the conclusion of this test.

Although the optimum output is obtained at a definite frequency, determined by the geometry of the elements inside the envelope, it has been observed that the external filament line may be used to shift the operating frequency over a bandwidth of 30 megacycles between half-power points. This feature is of considerable importance in allowing some tolerance in manufacturing, and in adapting the tube to several different types of radio-frequency circuits. It may be said that ordinary vacuum-tube production tolerances may be used to obtain tubes (from several different manufacturers) which all peak within a few megacycles of the same frequency.

The tube may be pulsed by biasing the grid beyond cutoff and applying suitable pulses to drive the grid positive several hundred volts for the required pulse interval. In another method of pulsing, the grid may be connected to the common radio-frequency point of the filament circuit through an appropriate resistor, and the filaments pulsed negatively with respect to the plate by as much as 30 kilovolts. Although efficiencies as high as 40 per cent have been observed for experimental tubes, the oscillator efficiency for production tubes is about 25 per cent under optimum conditions, and is constant for applied voltages greater than several kilovolts. This is interpreted as meaning that transit-time effects are negligible above this voltage, and rough calculations show that this should be the case. Frequency stability has been checked only during pulse operation, and within the uncertainty due to the Fourier components of the pulse, no instability could be found, nor could any frequency modulation be detected. An interesting feature of the tube characteristics is that the anode high-voltage filament power and blower can all be simultaneously turned on and off without any previous warm-up. No bad effects due to this practice have been observed.

V. Theory

Theoretical investigation of the properties of this type of design has proved to be difficult, primarily because of uncertainty about the current distribution in the closely spaced plate and grid loops at the frequencies mentioned. The general type of circuit is not new; similar circuits having been treated in papers by Holburn,[2] Mesny,[3] Gutton and Pierret,[4] and Denhardt,[5] and summarized in a paper by Wenstrom.[6] All of these treatments are concerned essentially with an experimental determination of the circuit characteristics.

Since in pulse operation the electron transit time is not appreciable, the tube probably should be capable of running with an efficiency of at least 50 per cent. Accordingly, the conventional-oscillator-design principles advanced by Prince about 1923 were applied to this tube by Lewis Greenwald, of this laboratory.

[2] F. Holborn, *Zeit. für Phys.*, vol. 6, pp. 328, 1921.
[3] Mesny, *L'Onde Electrique*, vol. 3, pp. 25–37; January, 1924.
[4] C. Gutton and E. Pierret, *L'Onde Electrique*, vol. 4, p. 387; 1925.
[5] A. Denhardt, *Zeits. für Hochfrequenz.*, vol. 35, pp. 212–223; June, 1930.
[6] W. H. Wenstrom, "An experimental study of regenerative ultra-short-wave oscillators," Proc. I.R.E., vol. 20, pp. 113–131; January, 1932.

These principles have been found to be quite adequate for ultra-high-frequency tubes so long as the transit time is small. The interloop coupling problem was avoided by considering the two loops and external load as a four-terminal network connected to the grids and plates, as shown in Fig. 3. It was assumed that this network maintained approximately 180 degrees phase difference between grid and plate and insured push-pull operation. On this basis, approximate explanations of many observed phenomena were made possible, and some improvement in efficiency was effected by increasing the plate-filament capacitance. However, the efficiency was not made to approach the value of 50 per cent. This may be due to use of incorrect effective values of interelement capacitances or to lack of sufficient information about the operation of the loops.

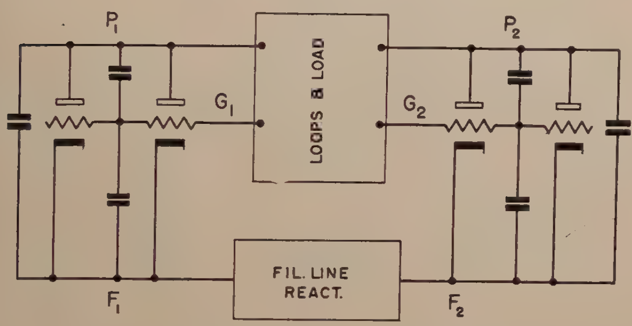

Fig. 3—Equivalent circuit of transmitting tube.

VI. Grounded-Plate, -Grid, and -Cathode Designs

It occurred to the authors, and to McCullough and Eitel in an unpublished communication, that the design herein described might be simplified by altering the internal construction so that the two anodes were tied together to have zero radio-frequency voltage difference (grounded plate). This would eliminate one internal tuned circuit, and would necessitate taking the power out at the filament line. Such a tube was built and it was found that almost exactly half the normal radio-frequency power could be obtained in this way. Subsequent tests on tuned-plate, tuned-grid tubes also showed that only half the power could be obtained from the filament lines, and work on grounded-plate tubes was then dropped without further investigation.

The authors attempted further designs involving grounded-grid and grounded-cathode construction, and the results were uniformly most unsuccessful, in that these types of tubes were never observed to oscillate. The grounded-grid construction was carried so far as to involve slits in the anodes to allow direct connection of the four grid cages by the shortest possible leads. The grounded-cathode tube was built around a rectangular box-type oxide cathode mounted between the two sides of the plate loop. In all cases, the element geometry was such as to have about the same amplification factor and transit time as for the tuned-plate, tuned-grid type of tube.

VII. Conclusion

This tuned-plate tuned-grid type of vacuum tube and circuit was developed a few years ago when low radio-frequency-impedance glass-metal seals had not been widely applied to tubes of this power and frequency. The subsequent development of such seals and corresponding tube improvement, especially in development of tubes which could be connected intimately with many different types of radio-frequency circuits which jointly covered a much wider band of frequencies, has been of great practical importance both in the war effort and in postwar applications. However, it is desired to point out one advantage of the tube of special importance for military use. Those who have seen extremely high-power pulse-transmitter circuits using the new low-impedance glass-metal-seal tubes are invariably impressed with the large size, weight, and complexity of concentric-line plumbing, a most descriptive term. Also, high-power pulse magnetrons for these frequencies are extremely large and heavy. Compared with these, the present tube is shipped from the tube manufacturer as a complete transmitter. The combined tube and shield weigh only about two pounds, occupy a much smaller volume, and are much simpler to make and assemble than are the newer type tubes and external oscillating circuits.

Tube type VT-158 was used principally in the light weight, early warning radar set AN/TPS-3, in both the European and Pacific theaters of operation.[7]

[7] H. A. Zahl and J. W. Marchetti, "Radar on 50 centimeters," *Electronics*, vol. 19, pp. 98–104; January, 1946.

Three New Antenna Types and Their Applications*

ARMIG G. KANDOIAN†, SENIOR MEMBER, I.R.E.

Summary—**Three newly developed types of antennas are described. The radiation pattern of each is substantially omnidirectional in the horizontal plane. The first has vertical polarization, the second has horizontal polarization, and the third is elliptically or circularly polarized.**

Variations of the above types, bandwidth considerations, tuning range, advantages, and limitations of each type are discussed, as well as the use of these antennas singly or in directive arrays for high power gain. Application to very-high-frequency and ultra-high-frequency broadcast, television, and link communication is briefly considered.

Photographs of experimental models giving construction details are shown, as well as measured characteristics of the type of antennas under consideration. Installation schematic of a typical array for frequency-modulation broadcasting is also shown.

Introduction

THREE new types of antennas have been developed for use primarily in the very-high-frequency and ultra-high-frequency spectrums. The radiation pattern of each antenna is essentially equivalent to that of a dipole; that is, it may be represented by a solid of revolution determined by a rotating figure of eight.

Type I, the Discone antenna, is intended primarily for vertical polarization, and, like a vertical dipole, gives an omnidirectional pattern in the horizontal plane. A distinctive feature of this antenna is its simplicity of construction and feeding. Its most important characteristic is satisfactory operation over a very wide band of frequencies (several octaves) without a substantial change of either input impedance or radiation pattern.

This type of antenna has wide applications wherever extremely wide frequency ranges are encountered and simplicity of mechanical design and installation are required.

Type II, a coaxially fed loop antenna, is intended primarily for horizontal polarization. The radiation pattern is omnidirectional in the plane of the loop. The radiators forming the loop are metallically supported from the mast or other supporting structure. Further, both supports and radiators form part of the coaxial feeding circuit. No balanced lines are used anywhere in the circuit. The bandwidth is controllable, though in general much narrower than with the type I antenna. No stubs are necessary to obtain a match to any common type of coaxial feeder of 50, 70, or 100 ohms.

The most important feature of this antenna is its simplicity of mechanical design and construction, and the ease with which a larger number of antennas may be "stacked" for high-degree directivity in the vertical plane while retaining omnidirectional radiation in the horizontal plane.

Typical applications of this type of antenna are frequency-modulation broadcasting, television, and general communication.

Type III antenna is similar to type II, except that at the center of the loop and perpendicular to it a vertical radiator has been added. The radiation pattern of type III is essentially the same as that of types I or II; the free-space field intensity at all points has both horizontal and vertical components in equal amounts. The type III, being a combination of an "electric" dipole (vertical radiator) and a "magnetic" dipole (horizontal loop), might, therefore, be called an "electric-magnetic" dipole.

Equality of amplitude of the two polarization components is not necessary as any desired ratio of amplitude as well as phase relationship between the horizontally and vertically polarized fields may be obtained.

The most interesting application of this type of radiator is in high-directivity arrays or in illuminating a highly directive parabolic reflector or horn for general communication application. The presence of both vertical and horizontal components, it is felt, will be helpful in reducing fading. There is also a possible application of this type of antenna to very-high-frequency and ultra-high-frequency broadcasting, where the receiving dipole could then be oriented for optimum signal-to-noise ratio.

Type I—The Discone Antenna

The Discone antenna, as the name suggests, consists of a disk and a cone whose apex approaches and becomes common with the outer conductor of the coaxial feeder at its extremity. The center conductor terminates at the center of the disk which is perpendicular to the axis of the cone and the feeding transmission line. Fig. 1 is a schematic of the Discone antenna with a tabulation of typical dimensions. Fig. 2 shows a Discone antenna.

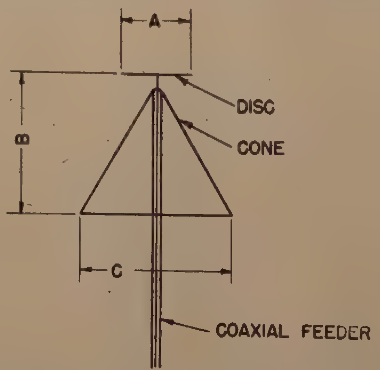

Fig. 1—Schematic diagram of the Discone antenna.

Cutoff Frequency	A	B	C
90 megacycles	18 inches	24 inches	20 inches
200 megacycles	9 inches	12½ inches	14 inches

The Discone antenna in its radiation behaves essentially as a vertical dipole. However, its change of impedance versus frequency is very much less than any

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† Federal Telephone and Radio Laboratories, New York, N. Y.

ordinary dipole of fixed length. The same is true of its radiation pattern.

From the circuit standpoint the Discone antenna is essentially a high-pass filter. It has an effective cutoff

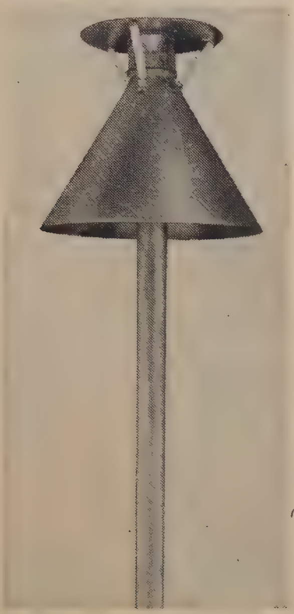


Fig. 2—Discone antenna for cutoff frequency of 200 megacycles.

frequency below which it becomes very inefficient, causing severe standing waves on the feeding coaxial line. Above the cutoff frequency, however, little mismatch exists and the radiation pattern remains substantially the same over a wide range of frequencies. The slant height of the cone is approximately equal to a quarter wavelength at the cutoff frequency.

Fig. 3 illustrates a typical mismatch versus frequency of a Discone antenna measured on a 50-ohm coaxial feeder.

Fig. 4 shows a typical measured radiation pattern of a Discone antenna, cut off at approximately 200 megacycles, measured every 50 megacycles up to 650. It is evident that, from cutoff (200 megacycles) all the way up to 650 megacycles, no drastic change in the radiation pattern has taken place. At the high-frequency end, however, the pattern does begin to turn upward. Investigation is now in progress to determine its characteristics at much higher frequencies.

The radiation patterns shown in Fig. 4 are measured patterns of the Discone antenna proper. As in most antennas with vertically polarized radiation, the supporting structure, in the present case the coaxial feeder, participates somewhat in the over-all radiation. This is due to currents induced in the supporting structure by the antenna. The amplitude of the induced currents in the present case is of the order of 5 per cent of the main antenna current and, hence, cause "scalloping" of the radiation pattern of approximately ± 5 per cent. For applications where it is required, a suppressing means, such as, for example, radial rods clamped on the supporting structure below the open end of the cone, may be used to prevent coupling between the antenna and the supporting structure.

An interesting application of the Discone antenna occurs when the cone of the antenna serves as the complete housing for the transmitting or receiving equipment.

A variation of the Discone antenna which has some useful applications is where the lower end of the cone is "grounded"; that is, made common with a large conducting surface.

Fig. 5 shows another experimental Discone antenna with a cutoff frequency of approximately 90 megacycles. Its performance is in general quite analogous to the small unit just described.

The Discone antenna may be visualized as a radiator intermediate between a conventional dipole and an electromagnetic horn. At the low end of its operating band, it behaves very much as a dipole; at much higher frequencies, it becomes essentially a horn radiator.

For information on antennas related to but not entirely equivalent to the Discone, the reader is referred to several investigators.[1–3]

Type II—Coaxially Fed Loop

Type II antenna is a loop radiator. This type is sometimes referred to as a "magnetic" dipole. An experimental model of the present type for approximately 550 megacycles is shown in Fig. 6.

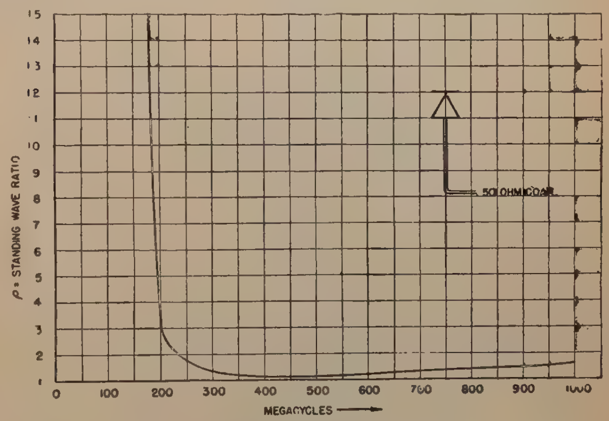


Fig. 3—Mismatch versus frequency of Discone antenna for a cutoff of 200 megacycles.

Loop antennas for very-high-frequency and ultra-high-frequency application have been described by several investigators.[4–7] While the present type of loop has the same radiation characteristic as an equivalent-size loop antenna previously described, its mechanical and electrical design is considerably different and offers some advantages and new possibilities in the application of the loop antennas.

1 S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Company, New York, N. Y., 1943, pp. 441–459.

2 G. C. Southworth, United States Patent Nos. 2,231,602 and 2,369,808.

3 W. L. Barrow, L. J. Chu, and J. J. Jansen, "Biconical electromagnetic horns," Proc. I.R.E., vol. 27, pp. 769–780; December, 1939.

4 Andrew Alford and A. G. Kandoian, "Ultra-high-frequency loop antennae," *Trans. A.I.E.E.*, (*Elec. Eng.*, 1940), vol. 59, pp. 843–848; 1940, and *Elec. Commun.*, vol. 18, pp. 255–265; April, 1940.

5 M. W. Scheldorf, "FM circular antenna," *Gen. Elec. Rev.*, vol. 46, pp. 163–170; March, 1943.

6 Jesse B. Sherman, "Circular loop antennas at ultra-high frequencies," Proc. I.R.E., vol. 32, pp. 534–538; September, 1944.

7 Donald Foster, "Loop antennas with uniform current," Proc. I.R.E., vol. 32, pp. 603–607; October, 1944.

Figs. 7, 8, and 9 show schematic diagrams of the type of loop under consideration. The metallic supporting structure as well as the radiators themselves form part of the coaxial feeding system. By proper choice of surge impedance of these two short lengths of line, in the supporting arm and the radiator itself, the desired impedance of 50, 70, or 100 ohms pure resistance may be obtained to match any common type of coaxial trans-

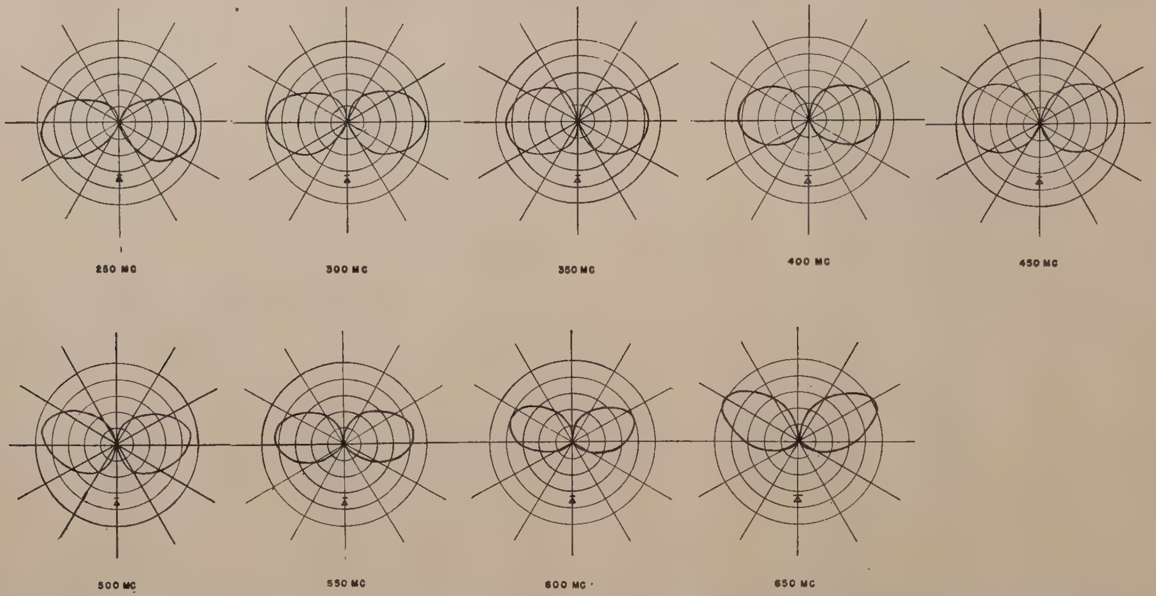


Fig. 4—Vertical radiation patterns of a Discone antenna for a cutoff frequency of 200 megacycles. Radiation patterns measured every 50 megacycles from 250 to 650 megacycles.

Fig. 5—Discone antenna for cutoff frequency of 90 megacycles.

mission line. Fig. 7 shows a single-element design in which the circumference of the loop is in the neighborhood of one-half wavelength or less. Figs. 8 and 9 illustrate multiple-element designs. No limitations, other than practical, exist for the number of elements making up the loop. Thus, a loop antenna of any diameter may be constructed and the current distribution maintained essentially uniform. The limitation of

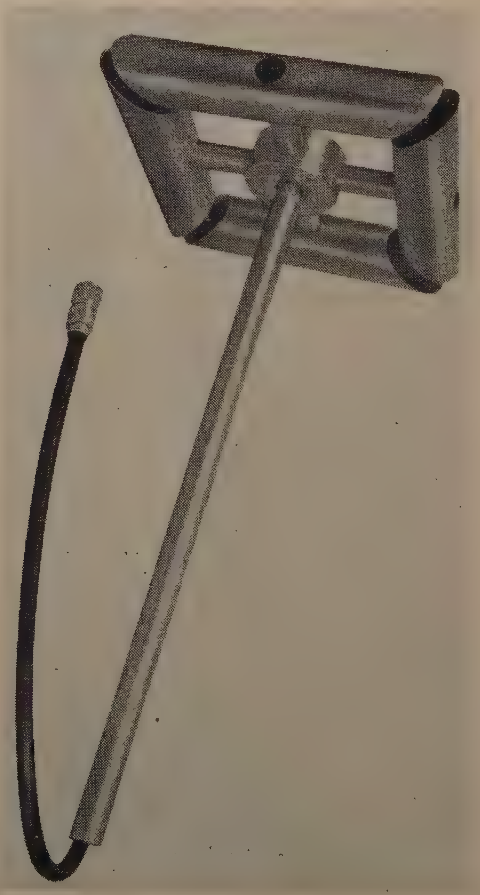


Fig. 6—Coaxially fed loop antenna for 550 megacycles.

diameter, small compared to a wavelength, need not be observed.

The radiation pattern of a loop with substantially uniform current distribution, and with a diameter small compared to a wavelength, is well known. In the plane of the loop the radiation is nondirective. In the plane perpendicular to the loop the radiation varies approximately as cos β, β being the angle measured from the plane of the loop.

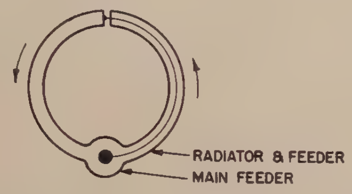


Fig. 7—Schematic diagram of single-element coaxially fed loop. Circumference approximately $\lambda/2$ or less.

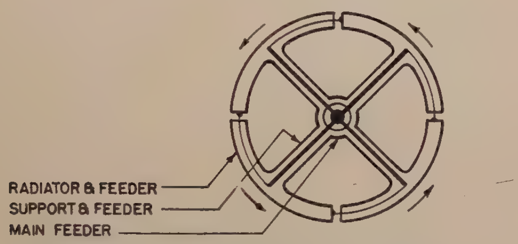


Fig. 8—Schematic diagram of 4-element coaxially fed loop. Circumference approximately 2λ or less.

Fig. 9—Schematic diagram of 6-element coaxially fed loop. Circumference approximately 3λ or less.

For loop antennas of any diameter and uniform current distribution, the radiation in the plane of the loop is still nondirective. In the plane perpendicular to the loop, Foster[7] has shown that the pattern is of the form $J_1(\pi d/\lambda \cos \beta)$ where d is the loop diameter to wavelength ratio and J_1 represents the Bessel function of the order unity.

An important consideration in any antenna is the input impedance. At very-high frequencies and ultra-high frequencies, the most practical way of expressing this information is in terms of standing waves on the line feeding the antenna. A typical measurement is shown in Fig. 10. The impedance at the center feed point is inductive below and capacitive above the midoperating frequency of 550 megacycles. The feeding line is a 50-ohm coaxial. Other experimental loops have been constructed giving both more and less bandwidth. In general, however, the bandwidth of a loop antenna cannot be made nearly so wide as a Discone type of antenna.

The loop type of antenna is particularly useful when high-degree directivity is desired in the vertical plane while retaining the omnidirectional pattern in the horizontal plane. This is accomplished by vertical stacking of any desired number of loops.

Fig. 11 depicts two experimental loops built as a pair and spaced approximately one wavelength. The input impedance to the pair is 50 ohms.

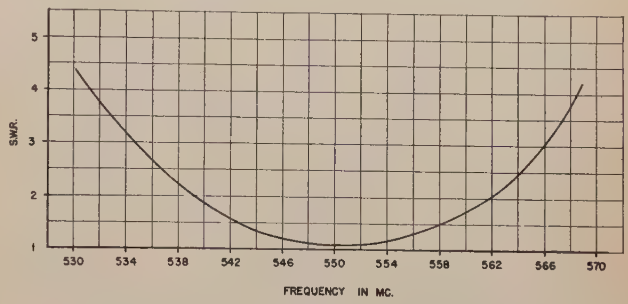


Fig. 10—Characteristic curve of frequency versus standing-wave ratio for a square loop. Four elements of 2-inch cross section; loop diameter, $8\frac{1}{2}$ inches.

Any number of such pairs may be stacked to give a desired amount of power gain. At the junction of the transmission line of any two pairs, however, a 2-to-1

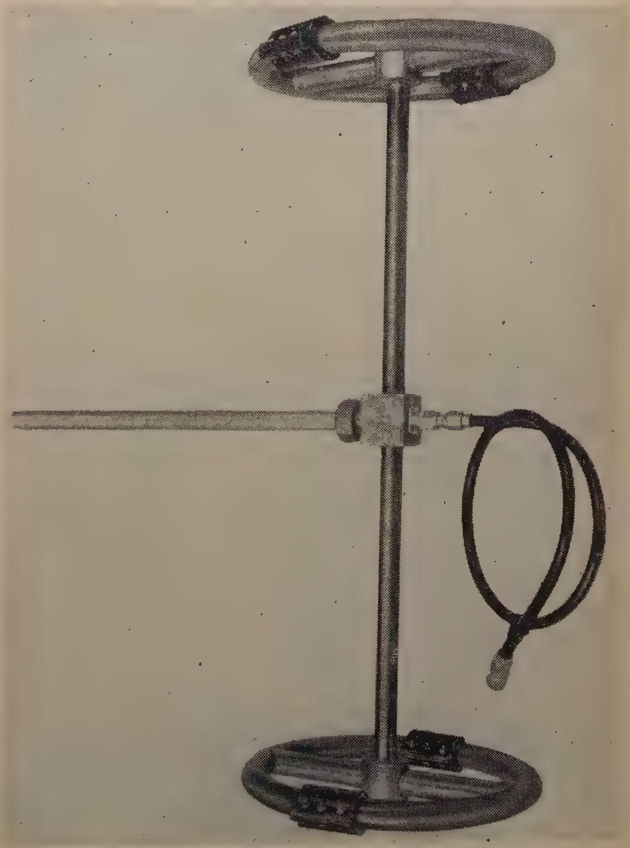


Fig. 11—Vertical stack of two loops with tuning stub.

impedance-correcting network is required to raise the impedance to the line impedance of 50 ohms (or any other desired level). A quarter-wave transformer probably is the most convenient network for this purpose. Fig. 12 shows the schematic of a feeding system for a stack of four loops for frequency-modulation broadcasting

Fig. 13 gives the necessary theoretical data to show what spacings to use between loops and what gains may be expected due to vertical stacking when equal currents are fed to successive loops.

A directive vertical pattern essentially free from minor lobes may be obtained by proper distribution of current between successive stacked loops.[8] However, under these conditions the over-all power gain from a fixed number of loops is reduced.

The characteristics and relative merits of the type of loop under consideration may be summarized as follows:

(a) No balanced feeders are used.

(b) No stubs are needed for matching; hence, the full bandwidth capability of the loop may be realized. A stub may be found desirable, however, to tune the loop, or a group of loops, over a wide frequency range.

(c) No insulator mechanical supports are necessary. Metallic supports are used, rigidly fastened to the mast and radiating members.

(d) Any size loop may be built, with essentially uniform clockwise or counterclockwise current distribution. Loop diameters of several wavelengths are feasible, and for certain applications desirable.

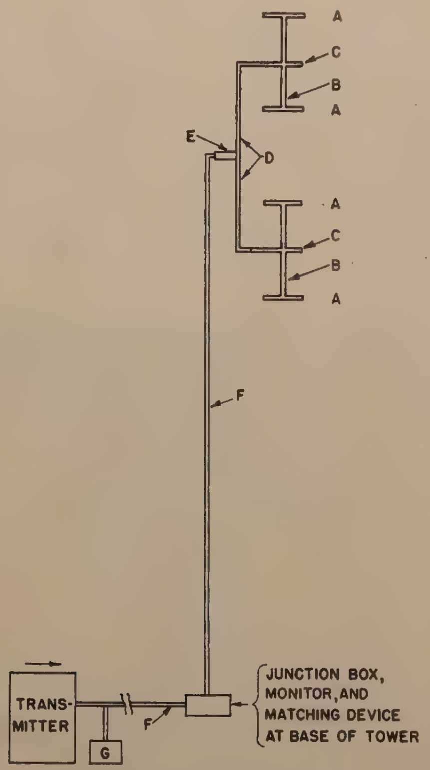


Fig. 12—Transmission line and feeding system for frequency-modulation broadcast antenna.
A = loop antennas
B = 100-ohm coaxial line
C = matching stub to be for operating frequency
D = 50-ohm line
E = quarter-wave 35-ohm line
F = 50-ohm coaxial line
G = pressurizing and air-drying equipment

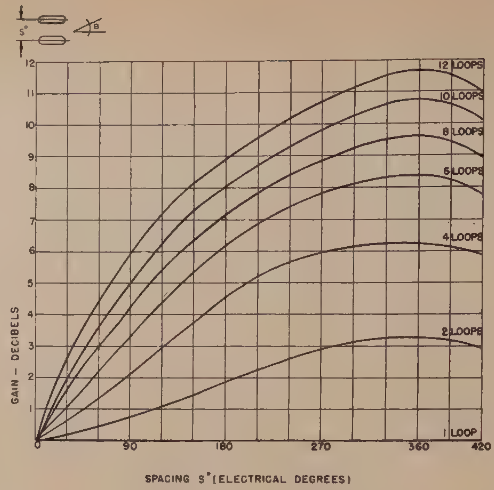
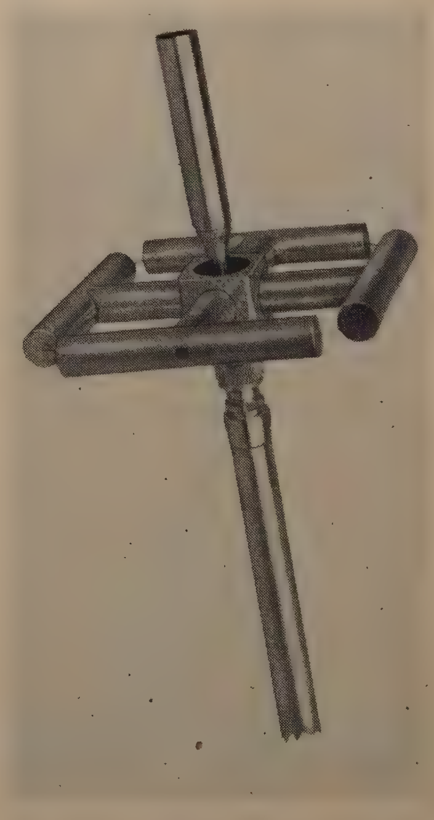


Fig. 13—Gain of linear array of loops vertically stacked.

$$(\beta) = \frac{\sin\left(\frac{ns^\circ}{2}\sin\beta\right)}{\sin\left(\frac{s^\circ}{2}\sin\beta\right)}\cos\beta$$

n = number of loops

$$\text{gain } (db) = 10\log_{10}\frac{1}{\frac{1}{n}+\frac{3}{n^2}\sum_{k=1}^{n-1}(n-k)\left[-\frac{2\cos ks^\circ}{(ks^\circ)^2}+\frac{2\sin ks^\circ}{(ks^\circ)^3}\right]}.$$



14—Type III antenna for an operating frequency of 1200 megacycles.

Type III—"Electric-Magnetic" Dipole

An experimental model of this type of antenna is illustrated in Fig. 14. This particular unit was built for use in the neighborhood of 1200 megacycles.

The "electric-magnetic" dipole consists essentially of a loop radiator of the type previously described, except

[8] A. G. Kandoian, "Ultra-high-frequency technique; radiating systems and wave propagation," *Electronics*, vol. 15, pp. 39–44; April, 1942.

that at the transmission-line junction at the center of the loop, and perpendicular to the plane of the loop, a vertical radiator is added. In effect, it is a combination of an "electric" and a "magnetic" dipole.

The radiation pattern of a horizontal loop, with a diameter of the order of $\lambda/2$ or less, is substantially the same as that of a vertical dipole. The only difference in the radiated field is in polarization. Thus, at every point in space equal amounts of horizontal and vertical polarization will be obtained, if the available power is equally divided between the loop and the dipole.

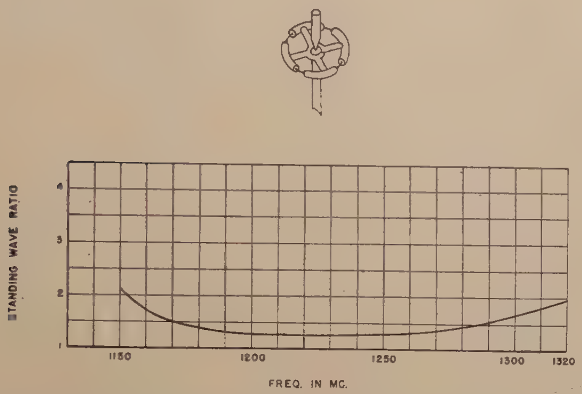


Fig. 15—Curve showing the standing-wave ratio versus frequency for the loop-dipole combination shown at top.

Control of the relative phase and amplitude of currents in the dipole and the loop is, of course, possible; and hence, any type of polarization desired can be produced; i.e., horizontal, vertical, circular, or the most general case, elliptical.

It is not difficult to demonstrate experimentally that at any point around such a transmitting antenna one can achieve a field strength independent of the orientation of the receiving dipole, provided that the receiving dipole is kept perpendicular to the direction of propagation.

In Fig. 15 is shown the mismatch-versus-frequency characteristic of an antenna of this type for operation in the neighborhood of 1200 megacycles. Fig. 16 shows another such experimental antenna for use around 350 megacycles.

In ultra-high-frequency communication networks where considerable fading may exist due to changes in the medium of propagation, this type of antenna will probably prove very useful. In cases of severe fading the probabilities are that both horizontal and vertical components of the electric field will not vary at the same rate and at the same time, since they are affected differently by the reflecting medium between the transmitter and receiver. Considerable improvement, therefore, should be experienced in reducing the over-all fading by the use of such a radiator, or a combination of such radiators in an array.

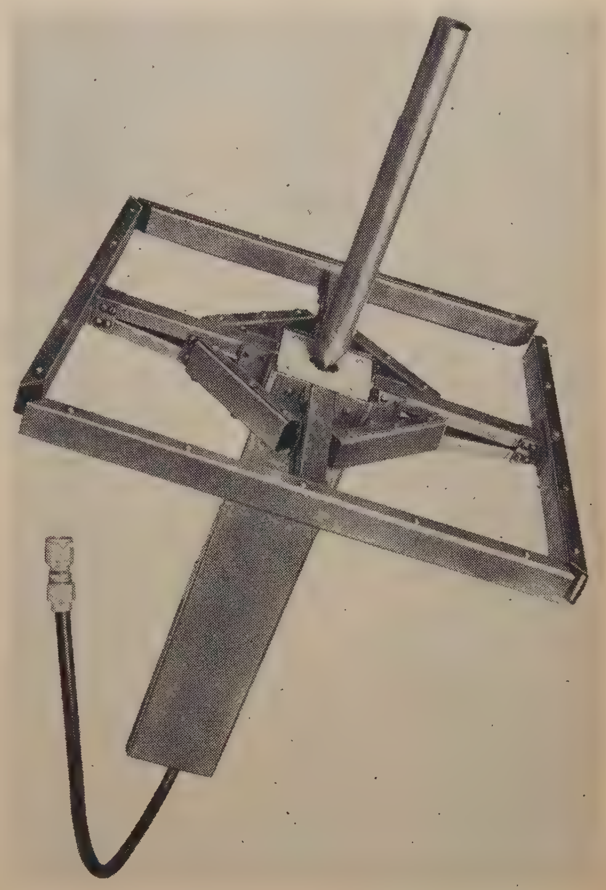


Fig. 16—Type III antenna for an operating frequency of 350 megacycles.

In addition, this type of radiator is particularly well suited for illuminating a large reflecting surface such as a paraboloid in a highly directive antenna system.

A possible important application of the "electric-magnetic" dipole is in the field of very-high-frequency and ultra-high-frequency broadcasting where the relative merits of horizontal as against vertical polarization have been under discussion for some years. If circular polarization were used at the transmitting end, it would permit the use of either a vertical or horizontal dipole at the receiving end, depending on convenience or architectural acceptability. The more particular listener would, of course, tend to orient his dipole to obtain the best signal-to-interference ratio, if there is any interference.

Whether any substantial improvement of performance would result in over-all reception can be determined only after field tests.

Decibel Conversion Chart*

ROBERT C. MIEDKE†, ASSOCIATE, I.R.E.

Summary—A decibel conversion chart has been designed for versatility and simplicity of use. This chart gives decibels directly from any two values of voltage, current, or power. It has two ranges; the lower range (scales *A*, *C*, and *E* of Fig. 1) are for voltage, current, or power ratios up to 10 to 1 and the extended range (scales *B*, *D* and *F*) are for voltage, current, or power ratios up to 10^6 to 1.

SEVERAL examples explaining the use of the decibel conversion chart are as follows:

(1) Assume a voltage ratio of 2.4 to 1.2. This voltage ratio in decibels is found by drawing a line from 1.2 on scale *A* to 2.4 on scale *E* passing through the decibel voltage scale *C* at 6 decibels. This can also be

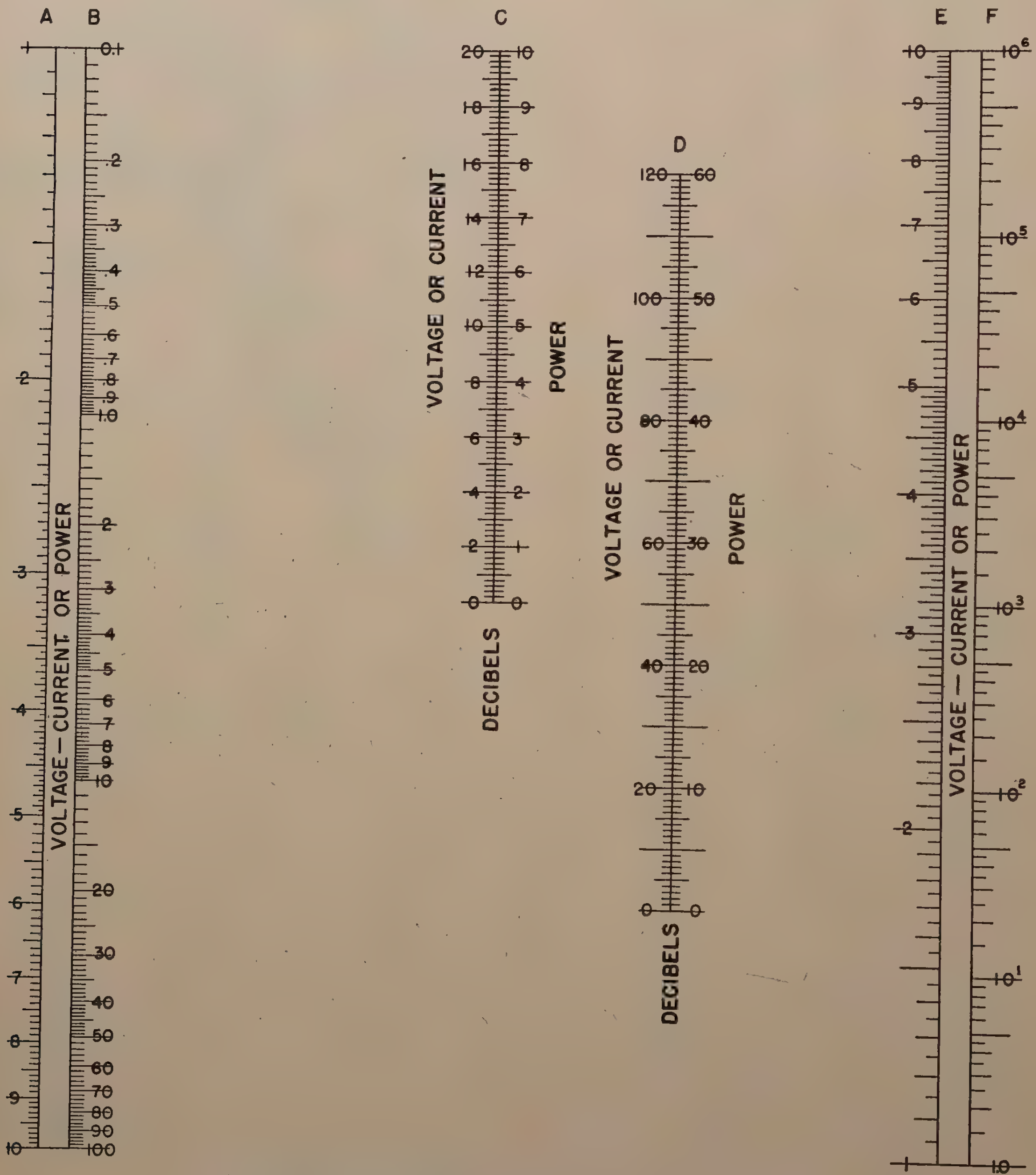


Fig. 1—Chart for converting current or power ratios to decibels.

* Decimal classification: R084. Original manuscript received by the Institute, September 24, 1945; revised manuscript received, November 5, 1945.

† Naval Research Laboratory, Washington, D. C.

found on scales B, D, and F; however, the decibel scale D is not expanded so much as the decibel scale C and cannot be read as accurately.

(2) Assume a voltage ratio of 1200 to 1.2. This voltage ratio in decibels is found by drawing a line from 1.2 on scale B to 1200 on scale F passing through the decibel scale D at 60 decibels.

(3) Assume a power ratio of 580 to 320. This power ratio in decibels is found by drawing a line from 3.2 on scale A to 5.8 on scale E (the ratio of 580 to 320 is the same as 5.8 to 3.2) passing through the decibel scale D at 2.6 decibels.

From the above examples it will be noted that the smaller value of any ratio will always be located on scale A or B and the larger value will always be located on the respective scale E or F. Also, the decibels corresponding to ratios of less than 10 to 1 can be found on either set of scales; however, it is preferable to use scales A, C, and E as the decibel scale C is expanded and can be read with greater accuracy.

Allied Military Government and Allied Commission in Italy

An interesting 125-page report entitled: "Review of Allied Military Government and of the Allied Commission in Italy" has reached the Institute from Italy through the courtesy of Rear Admiral Ellery W. Stone, U. S. Naval Reserve (A'14–M'16–F'24), Deputy President and Chief Commissioner of the Allied Commission since June 22, 1944. The magnitude of the tasks assumed in the temporary control of Italian affairs by the Allied Forces is clearly set forth in historical fashion in the report. The Institute is gratified that one of its leading members has played so constructive a part in a project of such magnitude.

Admiral Stone received the Award of Commander of the British Empire in March, 1945, and in December, 1945, he received the appointment as Knight Commander of the British Empire, the K. B. E.

Nontechnical Radar Broadcast

One of the first network broadcasts intended to explain for nontechnical listeners the principles and objectives of radar was given on September 6, 1945, in New York City when E. W. Engstrom (A'25, M'38, F'40) Irving Wolff (A'27–F'42), and J. E. Evans of RCA Laboratories and Stanley Parker of the U. S. Navy were interviewed by Miss Mary Margaret McBride. The fundamental principles, underlying the operation of radar and early experiments dating back to 1932, were covered by the RCA Laboratories representatives while Stanley Parker told about some of the actual experiences that he had had at sea with the operation of radar equipment. A short demonstration was given to show how a received radio wave when used in conjunction with a sharp microwave transmitting beam, could be used to show the direction of a reflecting object.

Cleveland Section

In *Cleveland Engineering*, published weekly by The Cleveland Technical Societies Council, the columns devoted to news of The Institute of Radio Engineers contain two items which are of interest to the entire membership of the I.R.E. Alfred J. Kres (A'39), secretary of the Cleveland Section, reported that a very small percentage of the questionnaire postcards, regarding proposed changes in the name of the Institute, had been returned.

The Cleveland Section subscribed 258 per cent over their quota in contributions to the Building Fund. A book containing the names of all contributors, intended to be kept in the lobby of the new building, went to press about December first.

Electronics in War and Peace

C. N. Kimball (A'34–M'40–SM'43), vice-president in charge of sales engineering for Aireon Manufacturing Corporation, recently spoke before the Advertising and Sales Executive Club of Kansas City on the use of early warning radar, ground control of fighter aircraft, blind bombing, and radar countermeasures in World War II. He directed his audience's attention to the proximity fuze and the recent acceleration of electronic research. The latter part of the speech was devoted to the carry-over value of these devices for peacetime functions, covering the navigational uses of radar and the part the absolute altimeter and terrain indicator will play in the safety of flight. Dr. Kimball mentioned the increase in speed of railway travel due to research in radio applications to railroad communication; industrial heating of plastics; uses of high-frequency radiation for cooking and heating applications; development of instruments to be used in the meat industry for determining humidity, frosting, and defrosting; and electronic instruments used in sorting, counting, measuring, oil prospecting, and color comparison.

It was pointed out that there exists an electronic solution or electronic aid for nearly every operation encountered in the course of manufacturing, packaging, or distributing, with the resultant increased safety, comfort, and leisure for the public.

Our New Home

BY AUTHORIZATION of the Board of Directors, The Institute of Radio Engineers has now purchased a building for its home. The Institute has acquired the Brokaw mansion on the northeast corner of Fifth Avenue and 79th Street in New York City. This building faces 79th Street, and is located in an extremely attractive residential neighborhood. It is only a block from the Metropolitan Museum of Art, which is diagonally across Fifth Avenue from the building.

The building is of chateau design, faced with gray granite and with steeply sloping tile roofs. It was constructed in 1889 for Mr. Brokaw, by Rose and Stone, architects. It has four stories, attic, basement, and cellar. The main body of the building is about 47×81 feet. It stands on a lot 51 feet 2 inches on Fifth Avenue and 110 feet along 79th Street. It faces south, and has a spacious yard on the easterly side of the building and open courtyards in front and on the Fifth Avenue sides, which spaces are in lawn. Since it does not occupy the whole lot, it is accessible to sunlight on three sides.

The building was purchased for $200,000 which is believed to be comparable with, or even less than, the value of the land. The mansions of the nineties and nineteen hundreds are being gradually sold. They are then torn down or converted to other uses, because of changes in living habits of the people who could afford to live in them. The property acquired by the Institute could be purchased for what is considered a justified amount because of a complicated ownership situation of a type which does not inspire real estate agents to devote much effort toward its sale, nor attract the speculator who is interested in acquiring property quickly at an opportune moment. Negotiations with three parties and court approval were necessary to complete the transaction.

The Board of Directors at its April, 1945, meeting instructed the Office Quarters Committee to begin negotiations toward the purchase of the property. The deeds were signed and the property

became ours on December 28, 1945.

The building from the exterior is one of the most attractive structures in that part of the city. It is well constructed throughout. Naturally, it will require some changes and alterations to suit it to the purposes of the Institute. However, these changes and alterations are materially less than would be necessary for many other buildings which the Committee seriously considered.

The Office Quarters Committee spent a year and a half searching for a structure considered suitable as a home for the Institute. At the start of this search there was available only a relatively small amount of money in surplus funds of the Institute, and the buildings available then within that limit were few and far between. None was considered a suitable permanent home for the Institute. Serious consideration was given to purchasing one as a temporary home because of the extremely pressing need for more space and the difficulty of finding rentable space in the city.

The Office Quarters Committee in its search inspected dozens of buildings of all kinds including ex-clubs, small office buildings, and many old mansions, and the members picked up on their faces, hands, and clothing considerable dust that may not have been disturbed since the eighteen nineties. After this search had progressed for well over six months, the Office Quarters Committee recommended that the Board approve a campaign to raise a building fund without waiting to have in mind a particular building that they desired to purchase. To have bound themselves to purchase a building before starting the drive for the fund would have involved risking a considerable portion of the Institute's reserves, and the Board wisely would not proceed on this basis. No "superbargain" was found that justified such a plan. It was believed that possible donors would be as willing to donate funds toward the purchase of a building to be selected later as to give funds to buy a particular building picked beforehand. In the latter half of 1944, therefore, the intensive search for a building was temporarily abated while plans were gotten under way to start the building-fund drive. Early in 1945 when the donations being received indicated that the goal desired likely would be reached, the Committee invited the Board to inspect the most promising buildings at prices falling within the range of the expected building fund. The Board was unhesitatingly in favor of the building which has just been purchased.

The Board and the Committee gave much attention and thought to the matter of acquiring a home which would be nearer the center of New York City, that is, nearer the railway stations, large hotels, and the three subway lines that tend to serve the center of Manhattan so thoroughly. Suitable and appropriate buildings, however, were very scarce and their price was very high. It would cost the Institute $100,000 to $200,000 more to acquire an equivalent home convenient to all the transportation centers than to purchase the building it has acquired. Neither the Committee nor the Board was persuaded that this extra expense was justified. The Institute staff will be the persons who are traveling to and from the building every working day. Such groups of long-term employees usually adjust their home locations to the site of their work. Committee Members in holding Committee meetings will travel to and from the building occasionally, and the Board will do so once each month or less frequently. Inquiry as to whether institutions located somewhat outside of the central business district find themselves handicapped by such locations led to the reply that they did not. People from outside of the city coming into the city usually do not find it any more difficult to reach a moderately outlying building by a taxicab than a more near-by one. Local commuting members who develop the habit of catching certain trains home are the ones who prefer always to be within a short distance of their railway terminal, and who particularly stressed their wish for what would be for them a more convenient location. The Committee and the Board, therefore, after considering all factors, felt they could not justify the purchase of a home in the highly expensive central business part of the city but that the location which they have secured will be satisfactory.

The building is a six-minute walk from the 77th Street stop on the Lexington Avenue subway and only about two more minutes' walk from the 86th Street stop, which is an express station. Fifth Avenue buses pass by the building on Fifth Avenue and there is a bus line running completely across 79th Street and passing the building. There seems to be a goodly supply of taxicabs passing up and down Fifth Avenue all the time.

The building has ample room for Committee meetings, Board meetings, and other normal activities. No attempt was made to find a building which could serve as a meeting place for a large group of members. The cost of such a building would not be justified as such meeting rooms would be used by the Institute only one or two days per month. Meetings of the New York Section will continue to be held either in the Engineering Societies Building on 39th Street or in other auditoriums in the city.

The Committee now is making detailed plans for alterations and decoration of the building for Institute purposes and to meet present safety regulations. It is not known when these alterations will be completed due to the difficult building-construction situation in the city. Every effort will be made to have the remodeling done as soon as possible in order that the staff, which is now divided between two locations, can be brought together again.

A photograph of the exterior of the building is to be found in this issue of the PROCEEDINGS. It is planned to publish in later issues the plans of the building, and photographs of parts of the interior. It is thought that the new headquarters of the Institute will be exceptionally attractive in appearance, and that particularly efficient operation of the work of the Institute can be carried out by the staff in the carefully planned new quarters. It is expected that the new building will have a home-like and welcoming atmosphere so that the visiting members will at last find commodious and cheerful I.R.E. quarters in which they can meet the staff and their fellow members and comfortably conduct whatever activities they have planned.

R. A. Heising, *Chairman*
Office Quarters Committee

Winter Technical Meeting

January 23, 24, 25, and 26, 1946

SUMMARIES OF TECHNICAL PAPERS

No papers are available in preprint or reprint form nor is there any assurance that any of them will be published in the PROCEEDINGS OF THE I.R.E. AND WAVES AND ELECTRONS, although it is hoped that many of them will appear in their pages.

CONSIDERATION OF FREQUENCY, POWER, AND MODULATION FOR A LONG-RANGE RADIO NAVIGATION SYSTEM

P. R. ADAMS
(Federal Telecommunication Laboratories, Inc., New York, N.Y.)

The general suitability of various types of radio transmission are examined with respect to the requirements of a long-range aerial-navigation system. By long-range navigation is meant, primarily, transoceanic navigation over distances of the order of 1500 miles.

The characteristics of short waves and long waves are separately and thoroughly analyzed with the help of considerable data which have been tabulated for this purpose. Data on other important factors are also considered, such as on signal field strength, in particular, for transmission below 300 kilocycles, also on atmospheric noise intensities, and on antenna radiating efficiency. A certain number of conclusions are obtained from these discussions. Modulation and other problems such as corona effect are also considered.

PHASE AND FREQUENCY MODULATION—A NEW METHOD

ROBERT ADLER
(Zenith Radio Corporation, Chicago, Ill.)
AND
F. M. BAILEY AND H. P. THOMAS
(General Electric Company, Syracuse, N. Y.)

The design of frequency-modulation transmitters has been simplified and their performance improved by the development of a new phase-modulator tube. In a concentric structure of conventional dimensions, a radial electron stream is shaped into a wave-like pattern which progresses continuously around the cathode. The development of the tube is reviewed. A description is also given of a commercial frequency-modulation broadcast transmitter making use of this tube together with a discussion of the design features involved.

ANTENNA FOR FREQUENCY-MODULATION STATION WGHF

ANDREW ALFORD
(Consulting Engineer, Cambridge, Mass.; formerly, Antenna and Direction-Finder Division, Radio Research Laboratory, Harvard University, Cambridge, Mass.)

An antenna of a new type is described. The radiation is horizontally polarized and essentially omnidirectional. One radiating element gives considerable gain over a dipole in the vertical plane. Design data and test results are discussed.

BROAD-BAND ANTENNAS AND DIRECTION-FINDING SYSTEMS FOR VERY HIGH FREQUENCIES

ANDREW ALFORD
(Consulting Engineer, Cambridge, Mass.; formerly, Antenna and Direction-Finder Division, Radio Research Laboratory, Harvard University, Cambridge, Mass.)
AND
J. D. KRAUS, A. DORNE, AND J. CHRISTENSEN
(Formerly, Antenna and Direction-Finder Division, Radio Research Laboratory, Harvard University, Cambridge, Mass.)

Methods have been developed to obtain broad-band antenna operation with low standing-wave ratio for frequency ranges employed in radar. The description of these methods includes directional antennas, antennas with circular polarization, slot antennas, nondirectional antennas, and also direction finders giving instantaneous visual presentation of direction for the same frequencies. These direction finders are usable over wide frequency ranges without any antenna adjustments and have a pickup sensitivity comparable to or exceeding that of a half-wave dipole. Examples of homing devices for use on airplanes are given.

DESIGN OF A SMALL SIZE HIGH-VOLTAGE RECTIFIER, TYPE 1Z2

GEORGE BAKER
(National Union Radio Corporation, Newark, N. J.)

The new factors in the design of a small-size, low-current rectifier tube for operation at several kilovolts are considered. The filament must withstand large mechanical forces produced by the electric field. The anode is shaped and processed to minimize cold field emission which, if uncontrolled, would result in large reverse current and high-velocity electrons striking the glass. The dielectric losses and leakage currents in the glass must be kept within limits.

MAGNETIC RECORDER AS AN ADJUNCT TO THE HOME RECEIVER

S. J. Begun
(Brush Development Company, Cleveland, Ohio)

BASIC PRINCIPLES OF UNDERWATER SOUND-EQUIPMENT DESIGN

Captain R. Bennett
(U. S. Navy, Bureau of Ships, Navy Department, Washington, D. C.)

The general subject of sound navigation and ranging is discussed in outline with some emphasis on the historical side of this little-known field.

The use of sonic and supersonic echo ranging and detection equipment for location of submarines is discussed briefly with special emphasis on the general principles of systems. Some of the special problems in this field of general interest are presented for information and possible study.

ELECTRONICS IN NAVAL WARFARE

Captain R. Bennett
(U. S. Navy, Bureau of Ships, Navy Department, Washington, D. C.)

A general survey of the use of electronic equipment in naval warfare is given with special emphasis on the progress made during the war. An attempt is made to show the interrelation of military uses of equipment with commercial uses. The effects of standardization and lack of it are indicated as they apply to the military and commercial fields. In this connection, some of the circuit developments made during the war are discussed briefly. This paper is nonmathematical.

NAVAL AIRBORNE RADAR

Captain L. V. Berkner
(U. S. Navy, Bureau of Aeronautics, Navy Department, Washington, D. C.)

Development of airborne microwave radar with improved indication gave enormous impetus to aircraft applications and produced advanced types of air to air interception, high- and low-altitude bombing, reconnaissance, submarine search, and many specialized applications. Several types of airborne radar and auxiliary devices are discussed and illustrated, including considerations of design and utilization problems. Limitations of advantages of airborne radar as a solution to future aircraft problems are briefly considered.

TEST EQUIPMENT AND TECHNIQUES FOR AIRBORNE RADAR FIELD MAINTENANCE

Captain E. A. Blasi and G. C. Schutz
(Radar Laboratory, Air Technical Service Command, Dayton, Ohio)

This paper discusses the various testing methods, techniques, and equipment that are used for the field maintenance of airborne radar systems. Techniques used in the measurement of frequency, power, and receiver sensitivity are outlined, as well as measurements of performance characteristics peculiar to airborne radar equipment. The specially designed instruments required for field maintenance and unique procedures devised to accomplish the measurement of radar performance are described.

FIELD INTENSITIES BEYOND LINE OF SIGHT AT 45.5 AND 91 MEGACYCLES

C. W. Carnahan, N. W. Aram, and E. F. Classen
(Zenith Radio Corporation, Chicago, Ill.)

This paper presents the results of a field-intensity monitoring project initiated by the Federal Communications Commission during the summer of 1945. Signal strengths from WMFM, a frequency-modulation broadcast transmitter at 45.5 megacycles, and an experimental 91-megacycle transmitter, both located at Richfield, Wisconsin, were continuously monitored for a period of two months at Deerfield, Illinois, a distance of about 80 miles. The data are analyzed in terms of the average field strengths at the two frequencies, and individual characteristics, such as the range and prevalence of fading and diurnal variation of average field strength. Comparison is made with field strengths predicted from Federal Communications Commission curves, and the ratio of the 91-megacycle field strength to that at 45.5 megacycles is found to be considerably lower than expected.

TWO NEW MINIATURE TUBES FOR FREQUENCY-MODULATION CONVERSION

R. M. Cohen, R. C. Fortin, and C. M. Morris
(RCA Victor Division, Harrison, N. J.)

The use of a new miniature converter tube and a new miniature radio-frequency amplifier tube in the headend unit of a frequency-modulation receiver covering the 88- to 108-megacycle band is described. The unit employs a radio-frequency stage, a converter stage, and two intermediate-frequency stages. Pertinent data covering construction and performance including circuit constants, stage gain, over-all gain, signal-to-noise ratios, image rejection, and oscillator-frequency drift are presented.

MICROWAVE MAGNETRONS

George Collins
(Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.)

A brief historical introduction reviews the wartime development and applications of microwave magnetrons. A review of the principles of operation and basic theories, as far as they are known, is presented, and a consideration of design principles, such as voltage and wavelength scaling and cathode problems, follows. Slides illustrating the construction of magnetrons designed for specific purposes are shown and actual operating conditions and characteristics given. A review of certain important characteristics of microwave magnetrons, both advantageous and otherwise, proceed some concluding remarks on special continuous-wave magnetrons suitable for communication purposes.

DESIGN OF COMMUNICATION RECEIVERS FOR THE NAVAL SERVICE WITH PARTICULAR CONSIDERATION TO THE VERY-HIGH-FREQUENCY AND ULTRA-HIGH-FREQUENCY RANGES

T. McL. Davis
(Naval Research Laboratory, Washington, D. C.)

Consideration is given to the unusual conditions under which naval communication receivers must function, the ideal receiver to meet these conditions, and the compromises which have been accepted in the interest of economy of space and weight, simplicity of operation and maintenance, and others which must be accepted due to limitations imposed by tube and circuit characteristics at the present state of developmental progress. The influence upon design as dictated by allocated channel spacing is discussed. The consideration is directed chiefly to the 100- to 400-megacycle band.

DEVELOPMENT IN RADIO SKY-WAVE PROPAGATION RESEARCH AND APPLICATIONS DURING THE WAR

J. H. Dellinger and Newbern Smith
(National Bureau of Standards, Washington, D. C.)

During the war, there were both a great and continuing improvement in world-wide coverage of ionospheric observing stations and development of rapid, simplified techniques for applying ionospheric data to military operational problems. The major aspects of the world-wide radio-propagation program, developed during the war, are described in this paper, some of which are geomagnetic "longitude effect" in the ionosphere, world-wide study of atmospheric radio noise, methods for calculating sky-wave field intensities, ionospheric-storm forecasting, effects of ionosphere on direction-finder indication, new frequency-allocation and -selection methods, and rapid dissemination of predictions and forecasts to the Allied Armed Services.

NAVY RADIO AND ELECTRONICS DURING WORLD WAR II

Commodore Jennings B. Dow
(U. S. Navy, Bureau of Ships, Navy Department, Washington, D. C.)

A decisive factor in our victories over the Axis Powers in World War II was the superiority of our radio-communication systems and the electronic equipment used in radar, sonar, loran, fire control, countermeasures, and ordnance. The paper deals with the vast expansion of the electronic industry to supply needed equipment, the organization of the Navy for research and development, procurement, installation and maintenance of the equipment. Statistical data are introduced to show the magnitude of the electronic-material problem, and the effort required to equip our vast fleet and its air arm with modern electronic apparatus in all categories.

GENERATION OF CONTINUOUS-WAVE POWER AT VERY HIGH FREQUENCIES

W. G. Dow, J. N. Dyer, W. W. Salisbury, and E. A. Yunker
(Radio Research Laboratory, Harvard University, Cambridge, Mass.)

Techniques for the generation of continuous-wave power at high frequencies have advanced greatly in recent years, both with respect to the amount of power that can be generated and the frequencies that can be reached. This paper describes a number of oscillator and amplifier techniques for the generation of continuous-wave power in the range of frequencies used by radar. These power sources are tunable over considerable frequency ranges such as 1.5 to 1, and can be modulated over relatively wide frequency bands. Arrangements employed include open-wire resonant lines with conventional tubes, concentric oscillators employing parallel-plane triodes, resonatrons, and magnetrons. Powers range from 10 watts upward, depending on the tubes employed, and the frequencies. The art has been advanced sufficiently to permit generation of at least 1 kilowatt of continuous-wave power up to the lower frequencies used in microwave radar, and powers as great as 30 kilowatts can be obtained at somewhat lower frequencies.

MICROWAVE CONVERTERS

C. F. Edwards
(Bell Telephone Laboratories, Inc., New York, N. Y.)

Microwave converters using point-contact silicon rectifiers as the nonlinear element are discussed with particular emphasis on the design of the networks connecting the rectifier to the input and output terminals.

Several converters which have been developed during recent years for use at wavelength between 3 and 30 centimeters are described. Some of the effects of the impedance-frequency characteristic of the networks on the converter performance are discussed.

IMPROVED CATHODE-RAY TUBES WITH METAL-BACKED LUMINESCENT SCREENS

D. W. EPSTEIN AND L. PENSAK
(RCA Laboratories, Princeton, N. J.)

Considerably improved cathode-ray tubes result from the application of a thin metallic layer on the beam side of the luminescent screen. Observations and measurements of such tubes show the advantages of increased light output, improved contrast, elimination of secondary-emission difficulties, and, under appropriate conditions, the elimination of ion spot.

SOME TECHNICAL DEVELOPMENTS IN LIGHT-WAVE COMMUNICATIONS

COMMANDER J. M. FLUKE AND LIEUTENANT (j.g.) N. E. PORTER
(U. S. Naval Reserve, Bureau of Ships, Navy Department, Washington, D. C.)

The purpose of this paper is to present and describe some of the recent technical advances in methods and components utilized in light-wave-communication devices. Reasons for exploitation of communication possibilities in the so-called light-wave portion of the spectrum are initially discussed, and some comparison is made with other communication methods. Included in this discussion, also, are the features of some of the new types of light-energy radiators, filter developments, converter tubes, highly sensitive receiving elements, and the associated electronic power supply and amplifying equipment.

A MEDIUM-POWER TRIODE FOR FREQUENCIES ABOUT 600 MEGACYCLES

S. FRANKEL AND J. J. GLAUBER
(Federal Telecommunication Laboratories, Inc., New York, N. Y.)
AND
J. WALLENSTEIN
(Federal Telephone and Radio Corporation, Newark, N. J.)

A tube is described which was originally designed for pulse operation to deliver approximately 50 kilowatts peak power output at 600 megacycles with good efficiency.

Design considerations are discussed which include, as the most important factor, problems of transit time, peak emission, cooling, and circuit properties of the internal tube structure.

A detailed description of this tube structure is given which includes design considerations of the electrodes, operating conditions, and static characteristics. Uniformity of characteristics is also discussed.

Methods of testing and using the tube as an oscillator, amplifier and frequency multiplier are described and the results obtained are given. These results include data and curves of power output versus frequency; efficiency versus frequency; power gain as an amplifier and multiplier; and life-test results.

THE ROLE OF ATMOSPHERIC DUCTS IN THE PROPAGATION OF SHORT RADIO WAVES

J. E. FREEHAFER
(Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.)

Experience gained during the war has shown that strong fields, at frequencies greatly exceeding the penetration frequency of the ionosphere, are often observed at several times the horizon distance. Experimental and theoretical investigations in this country and in England indicate that these effects are associated with the presence of layers in the troposphere in which the vertical gradient of refractive index exceeds numerically the reciprocal of the earth's radius. Under these conditions a duct is former which for sufficiently high frequencies both reduces the rate at which the field is attenuated with range and also disturbs the normal height-gain effect.

A SPECTRUM ANALYZER FOR MICROWAVE PULSED OSCILLATORS

F. J. GAFFNEY
(PIB Products Company, Brooklyn, N. Y.; formerly Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.)

A spectrum analyzer is described which utilizes a sharply tuned receiver whose response frequency is swept across the frequency spectrum to be analyzed at a rate slow compared to the pulse recurrence rate. The receiver output pulses are displayed on the screen of a cathode-ray oscilloscope whose horizontal sweep is synchronized with the variation of receiver response frequency. The envelope of these pulses represents the Fourier transform of the pulsed oscillator output. The distortion produced by the finite bandwidth of the receiver is analyzed. Design details of the radio-frequency input plumbing are discussed.

ELECTRICAL CHARACTERISTICS OF QUARTZ-CRYSTAL UNITS AND THEIR MEASUREMENT

W. D. George, M. C. Selby, and R. Scolnik
(National Bureau of Standards, Washington, D. C.)

An investigation of methods of accurately measuring the performance characteristics of two-terminal quartz-crystal units is presented. Relative merits and limitations of these methods are listed and forms of presentation of performance characteristics are indicated. Some observations of crystal-unit behavior, most useful to the radio-and-electronic engineer, are given. These may be interpreted and predicted on the basis of the performance characteristic curves suggested. It is believed that a practical approach is described whereby relatively simple measuring apparatus may be used in studying and standardizing those constants and characteristics that specify the quartz-crystal unit in itself apart from any external circuits with which it may be applied.

TELEVISION IN THE ULTRA-HIGH FREQUENCIES

Peter C. Goldmark
(Columbia Broadcasting System, New York, N. Y.)

The Columbia Broadcasting System ultra-high-frequency television system and results of recent tests are described and considerations of color in television discussed.

MICROWAVE TRIODES ADAPTED TO MODERN USAGE

Everett M. Goodell
(Sylvania Electric Products, Inc., Emporium, Pa.)

A new series of planar-grid triodes has been developed which is adaptable both to pulse and continuous operation at microwave and lower frequencies. Special features incorporated into the mechanical design permit quick changes and adaptability to a wide variety of usage.

A NEW SYSTEM OF ANGULAR VELOCITY MODULATION EMPLOYING PULSE TECHNIQUES

James F. Gordon
(Bendix Radio Division, Baltimore, Md.)

A frequency-modulation system is described in which a crystal-controlled pulse triggers a multivibrator. This pulse establishes a reference time for the system. The asymmetry of the multivibrator cycle varies with modulation. Clipping the reference pulse and differentiating the intelligence pulse which is generated at the cross-over point of the multivibrator cycle produces a source of time-modulated intelligence. These pulses are used to control the phase of a continuous-wave carrier.

LINEAR SERVO THEORY

R. E. Graham
(Bell Telephone Laboratories, Inc., New York, N. Y.)

The servo system is presented as a feedback circuit. Typical components of electromechanical circuits are described in homogeneous terms by means of a conventional analogy. The nomenclature of frequency analysis is used to describe the servo circuit as a transmission system.

The problems of linear servo design are discussed in the language of electrical-circuit feedback theory. A simple logarithmic frequency plot is found adequate for most design considerations. Exact and approximate methods for calculating dynamic errors in follow-up systems are described, and a discussion is given of the interrelations between input signal, dynamic error, transient response, and noise vulnerability.

Linearization of motor-drive systems by use of local velocity feedback and pre-equalization of the input signal to reduce over-all error are described.

AIRBORNE RADAR EQUIPMENT FOR AIRCRAFT INTERCEPTION

Major F. L. Holloway, Captain R. P. Burrows, and J. E. Keto
(Radar Laboratory, Air Technical Service Command, Dayton, Ohio)

This paper describes some of the outstanding airborne radar equipments and systems developed during the war for the radar-bombing operations of the Army Air Forces. Technical characteristics and requirements of such equipment are presented. Problems introduced by the technical and operational aspects of radar bombing are discussed as well as the effect of such problems on the design of the subject equipment.

TUNABLE RECEIVERS FOR VERY HIGH FREQUENCIES

G. E. Hulstede, J. M. Pettit, H. E. Overacker, K. Spangenberg, and R. R. Buss
(Radio Research Laboratory, Harvard University, Cambridge, Mass.)

Developments have greatly advanced receiver techniques in the direction of obtaining tunability, good

sensitivity, and wide frequency coverage at increasing high frequencies. This paper describes techniques that have been employed in a line of receivers that give continuous coverage of 10,000 megacycles. The most advanced of these equipments have high image rejection even at microwave frequencies, have single-dial control, a sensitivity reasonably close to the theoretical ultimate, and are tunable over frequency ranges of approximately 2 to 1 or more.

THE THEORY AND APPLICATION OF THE RADAR BEACON

CAPTAIN R. D. HULTGREN AND L. B. HALLMAN, JR.
(Watson Laboratories, Red Bank, N. J.)

Part I discusses the general theory underlying the operation of radar beacons. The various components of a typical beacon, such as the receiver, discriminator, modulator, coder, and transmitter, are discussed in some detail. System considerations such as factors governing choice of operating frequencies, required receiver sensitivity and transmitter power, choice of pulse duration, the cause and effect of delay in the beacon, and type of coding, are discussed. Part II discusses the application of radar beacons to aircraft homing, landing, rendezvous, identification, and airport surveillance. Special types of radar beacon applications such as the Beacon Blind Approach System (BABS), sea rescue devices, and communication aids are discussed in some detail.

THREE- AND NINE-CENTIMETER PROPAGATION MEASUREMENTS IN LOW-LEVEL OCEAN DUCTS

M. KATZIN AND R. W. BAUCHMAN
(Naval Research Laboratory, Washington, D. C.)

In order to check the effect on three- and nine-centimeter transmissions of low-level ducts formed in oceanic air, one-way measurements between a ship and a shore station were made with antenna combinations of various heights. The experimental setup and techniques are described. Meteorological measurements taken from both shore and ship are described, and the variations in duct height and strength discussed. Meteorological and radio measurements inland were made, and the effect of distance back from the shore on meteorological conditions and radio transmission are described. Three-centimeter radar observations were made during the latter part of the project, and the variation of echo amplitude with range of ship target was measured. Analysis of radio and radar measurements is given.

NAVAL WARFARE COMMUNICATIONS PROBLEMS

COMMANDER J. O. KINERT
(U. S. Navy, Naval Operations, Navy Department, Washington, D. C.)

A brief review of problems encountered during the development and growth of amphibious assault techniques is given. How problems were met from point of view of electronic material with an evaluation of available types is discussed. Recommendation for future design is also given.

METAL-LENS ANTENNAS
(With Demonstration)

W. E. KOCK
(Bell Telephone Laboratories, Inc., New York, N. Y.)

A new type of antenna is described which utilizes the optical properties of radio waves. It consists of a number of conducting plates of proper shape and spacing and is, in effect, a lens whose focusing action is due to the high-phase velocity of a wave passing between the plates. Its field of usefulness extends from the very short waves up to wavelengths of perhaps five meters or more. The paper discusses the properties of this antenna, methods of construction, and applications.

TWO MULTICHANNEL MICROWAVE RADIO-RELAY EQUIPMENTS FOR THE U. S. ARMY COMMUNICATION NETWORK

RAYMOND E. LACY
(Coles Signal Laboratory, Red Bank, N. J.)

Radio set AN/TRC-5 and radio set AN/TRC-6 are described whereby the U. S. Army Signal Corps has pioneered in applying the art of microwave techniques to communication equipments in order to provide transportable multichannel radio-relay sets for interconnection with high-grade voice circuits. An explanation is given of radar-type pulse-time-division methods of modulation which eliminate the relatively cumbersome carrier-frequency terminal equipment usually associated with a multiplex system. Audio performance and radiation characteristics are shown which presage a microwave epoch for long-lines communication.

CAPACITANCE-COUPLED WIDE-BAND INTERMEDIATE-FREQUENCY AMPLIFIERS

MERWIN J. LARSEN AND LYNN L. MERRILL
(Stromberg-Carlson Company, Rochester, N. Y.)

The design and performance of capacitance-coupled intermediate-frequency amplifiers is discussed, with

particular emphasis on applications to television receivers. Experimentally obtained curves are presented as an aid in selecting the circuit parameters necessary to produce flat-top frequency response for various operating ranges. The design of capacitance-coupled traps is treated theoretically, and experimental evidence is presented to show the validity of the theoretically predicted results.

A THREE-BEAM OSCILLOGRAPH FOR RECORDING AT FREQUENCIES UP TO 10,000 MEGACYCLES

Gordon M. Lee
(Central Research Laboratories, Red Wing, Minn.)

Transit-time distortion is a fundamental limiting factor in the application of high-speed cathode-ray oscillographs to the recording of high-frequency voltages or fast transients. A three-beam, high-speed, micro-oscillograph is described in which the transit-time reduction in deflection sensitivity is calculated to be but 4 per cent at 3000 megacycles and 40 per cent at 10,000 megacycles. Single-sweep oscillograms of 3000 and 10,000 megacycle oscillations and 10^{-9}-second transients are shown.

CASCADE AMPLIFIER KLYSTRONS

E. C. Levinthal
(Sperry Gyroscope Company, Inc., Garden City, L. I., N. Y.)

The theory of operation of a three-resonator klystron amplifier is discussed for the case of small-signal operation. An analogy between the operation of a cascade amplifier and velocity modulation by a saw-tooth voltage is demonstrated. Some calculated values of tube parameters are given which yield theoretical efficiencies in the neighborhood of 74 per cent. Data are presented which verify the production of greater efficiency for large-signal operation.

ENEMY RADIO AND RADAR EQUIPMENT

Lieutenant Commander E. L. Luke and John C. Link
(Naval Research Laboratory, Washington, D. C.)

ULTRA-HIGH-FREQUENCY TELEVISION RECEIVERS

Harold T. Lyman
(Columbia Broadcasting System, New York, N. Y.)

Combined black-and-white and color television receivers, both in direct-view and projection types, have been developed in the Columbia Broadcasting System laboratories. These receivers operate in the television band between 480 and 920 megacycles with sight and sound on one carrier. Details of circuit design, mechanical layout, and performance data are presented.

APPLICATION OF RADAR TECHNIQUES TO AIRCRAFT FIRE-CONTROL SYSTEMS

Major E. A. Massa, Captain I. Paganelli, and Captain Fred A. Best, Jr.
(Radar Laboratory, Air Technical Service Command, Dayton, Ohio)

This paper will discuss the basic problems involved in aircraft fire-control systems and the three basic types of radar systems which have been developed to solve these problems. A more detailed description of several of these equipments, including equipments which depart from conventional radar design, will be presented. The advantages which have been gained by the use of radar in this field will be listed and analyzed. A short discussion of future radar fire-control systems concludes the paper.

THE ROLE OF ELECTRONICS IN ANTIAIRCRAFT GUN-FIRE CONTROL

Captain F. B. MacLaren
(Ordnance Department, Frankford Arsenal, Philadelphia, Pa.)

During the war, our major caliber antiaircraft equipment advanced from an essentially mechanical system to one employing electronics as a vital part. A detailed description of the circuits involved in associated radar, computer, servomechanism, and proximity-fuze components, compared with equivalent enemy material, will explain the high degree of effectiveness obtained in combat. However, many improvements are still needed in electronic systems to supplement the defense against weapons such as the German V-2 rocket, especially if augmented with atomic power.

SECONDARY-EMISSION CATHODES FOR MAGNETRONS

J. W. McNall, H. L. Steele, Jr., and C. L. Shackelford
(Westinghouse Electric Corporation, Bloomfield, N. J.)

A few weeks before Pearl Harbor, an investigation was begun at the Massachusetts Institute of Technology Radiation Laboratory on cathodes for microwave multicavity magnetrons. The cathode used for such tubes was an indirectly heated oxide-coated cathode. It was known that the cathode temperature increased considerably due to back-bombardment by electrons when the magnetron was operated, and, in fact, it was believed that most of the emission current was not of thermionic but rather of secondary origin. Realizing that the electron back-bombardment of the cathode would impose undesirable limitations on the duty cycle and maximum average power at which magnetrons could be operated, it was decided to investigate the feasibility of using a cold secondary-emitting cathode in place of the thermionic one generally used. Since bombardment of the cathode can occur if the magnetron is oscillating, and since the magnetron will oscillate only

if there are electrons available in the cathode-anode space, it is obvious that an auxiliary primary emitter must be provided. This emitter is required to supply, however, only enough current to initiate oscillations in the magnetron. At the Radiation Laboratory, many different materials were used as secondary emitters, including beryllium, aluminum, nickel-barium alloy, thorium, and silver-magnesium alloy.

At the beginning of 1943, this project was discontinued at the Radiation Laboratory and was taken up in the Westinghouse Lamp Division Research Department. The work was directed along several lines including investigations of activation processes, methods of cooling the secondary-emitting cathode, investigations of several more materials, testing at higher duty cycles, studies of the effects of heating, measurements of the secondary emission yield of the surface, and correlations of these yield measurements with the maximum peak emission currents obtainable for various voltages and states of activation.

FREQUENCY ALLOCATIONS

COMMANDER PAUL D. MILES
(Federal Communications Commission, Washington, D. C.)

A brief review of international and domestic frequency-allocation regulation is given, followed by a description of the steps taken in the past two years by the United States to prepare for the next international telecommunications conference. Consideration is given to new radio services which must be anticipated, their probable effect upon future allocation tables, and improved techniques which must be exploited to insure the accommodation of these services in the spectrum.

PROBLEMS ASSOCIATED WITH THE STANDARDIZATION OF QUARTZ-CRYSTAL UNITS FOR MILITARY EQUIPMENT

CAPTAIN CHARLES J. MILLER, JR.
(Squier Signal Laboratory, Fort Monmouth, Red Bank, N. J.)

Some of the technical problems which must be solved before a series of ultimate standard quartz-crystal units for use in military signal equipment can be established are discussed, and progress to date in the solution of these problems is indicated. The crystal impedance meter, a new instrument for measuring the impedance of a crystal unit, is introduced and described. A brief review of the fundamental electrical properties of crystal units and associated oscillator circuits is included to furnish a technical background upon which to discuss these problems. A review of the three principal applications of crystal units in frequency-control circuits, calibrator circuits, and filter circuits, and a review of the various types of crystal holders and methods of mounting are included.

GLASS PROBLEMS IN THE MANUFACTURE OF MINIATURE TUBES

HENRY J. MILLER
(RCA Victor Division, Harrison, N. J.)

The use of miniature tubes in military equipment made it necessary to develop stem making and stem-to-bulb sealing techniques to provide tubes which would withstand without glass failure severe mechanical and thermal loading and shock as well as vibration. The three factors which govern the mechanical stability of the stem are buffer gas, depth of insertion, and strain. Methods of controlling these factors are described. The stem-to-bulb sealing technique required to obtain a seal of maximum mechanical rigidity is outlined. It is shown that the resulting procedures for the mass production of tubes were successful in withstanding the severe mechanical stress of wartime use without any epidemic glass failure in the field.

NOISE SPECTRUM OF CRYSTAL MIXERS

P. H. MILLER
(University of Pennsylvania, Philadelphia, Pa.)

Studies of the noise spectrum of crystal mixers were made over the frequency range of 50 to 1,000,000 cycles per second. In the audio range a band-pass filter employing a three-terminal Wien bridge was used to perform the analysis. For analysis at higher frequencies 7 megacycles were added to the noise spectrum and a communications receiver employed. It was found that regardless of the type of power applied to the crystal the noise temperature varies inversely with the frequency. The noise in the audio range is always large, a noise temperature ratio of 10^6 being typical. A mechanism responsible for the observed noise has not yet been suggested.

A NEW SYSTEM OF RADIO TELEMETERING

DAVID W. MOORE, JR., AND FRANK G. WILLEY
(Fairchild Camera and Instrument Corporation, Jamaica, L. I., N. Y.)

The Fairchild system of radio telemetering was designed to transmit indication of aircraft instruments from plane to ground over a conventional aircraft radio transmitting and receiving equipment. Ground indication is made by an indicator which closely simulates the standard aircraft instrument in appearance and modification to the aircraft instrument in the plane in flight. Instrument indication is transferred into electrical phase angle which may be transmitted by radio equipment. This phase angle is then compared with a reference signal in the receiving station and converted to a dial indication. Provision may be made for sending a number of instrument indications over a single carrier,

enabling transmission of indications of a flight group of instruments from, for example, a radio-controlled plane or missile to a distant operating point, either on the ground or in another plane.

MICROWAVE POWER MEASUREMENT

T. Moreno and O. C. Lundstrum
(Sperry Gyroscope Company, Inc., Garden City, L. I., N. Y.)

Possible methods of microwave power measurement are reviewed. The design requirements for bolometric wattmeters are outlined, and examples are given of bolometer elements that have been developed to meet these requirements. A recently developed bolometer element that may be used over an exceedingly wide band of frequencies is included. The results of experiments to determine the accuracy of these wattmeters are summarized; these experiments indicate that errors may be held to within a few per cent.

DIRECTIONAL COUPLERS

W. W. Mumford
(Bell Telephone Laboratories, New York, N. Y.)

The directional coupler is a device which samples separately the direct and the reflected waves in a transmission line. A simple theory of its operation is derived. Design data and operating characteristics for a typical unit are presented. Several applications which utilize the directional coupler are discussed.

AIRCRAFT AUTOMATIC POSITION PLOTTER

A. C. Omberg and W. L. Webb
(Bendix Radio Division, Baltimore, Md.)

The development of a device for feeding information from two automatic direction finders and a magnetic heading device into a computer, which changes polar co-ordinates into rectangular co-ordinates, is described.

The rectangular co-ordinates give the position of an aircraft with respect to two ground transmitters automatically controlling a "crab" which continuously plots the position of the aircraft on a chart. The method of developing the computer and a model of the device is described.

RADAR MODEL XAF

R. M. Page
(Naval Research Laboratory, Washington, D. C.)

This paper gives the story of the technical development of the U. S. Navy's first operational radar equipment, the XAF. Theoretical considerations and design parameters are given for pulse-modulation transmission and reception equipment, and exemplified in the XAF circuits. Of particular interest are the six-tube, high-frequency oscillator and the coupling circuits permitting efficient operation of both transmitter and receiver on the same antenna.

A NEW HIGH-SPEED RECORDING POTENTIOMETER

V. L. Parsegian
(Portable Products Corporation, C. J. Tagliabue Division, Brooklyn, N. Y.)

A new high-speed recording potentiometer is described, having a carriage travel of 10 inches in one second and balancing to within an accuracy of 0.1 per cent of full range. The instrument is of the photoelectric type, using a new sturdy, short-period, low-inertia-error galvanometer, a photocell, and a phase-reversing amplifier which drives a split-phase carriage motor when the photocell illumination exceeds, or is less than, an intermediate "balance" value. Means for canceling the galvanometer lag is described, as well as a new thyratron circuit for printing records.

AN INTRODUCTION TO HYPERBOLIC NAVIGATION

J. A. Pierce
(Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.)

In less than five years, loran, the American embodiment of the new method of navigation, has grown from a concept into a service used by tens of thousand of navigators over three tenths of the surface of the earth. The first part of the present paper describes the history of this program. A second section deals with the fundamental concepts of hyperbolic navigation and gives some details regarding the kinds of equipment now employed for transmission, reception, and interpretation of pulse signals for this service.

JOINT ELECTRON TUBE ENGINEERING COUNCIL

O. W. Pike
(Joint Electron Tube Engineering Council)

This paper outlines the conditions leading up to the formation of the Joint Electron Tube Engineering Council as the joint body responsible for the technical aspects of tube standardization in the National Electrical Manufacturers Association and the Radio Manufacturers Association. The basic organization of JETEC is described, and its relation to the parent organizations and other agencies such as the Joint Army-Navy Committees is outlined. The activities of JETEC during the past year are summarized and its program for the future indicated.

MAGNETRON CATHODES

M. A. Pomerantz
(Bartol Research Foundation of the Franklin Institute, Swarthmore, Pa.)

This paper will include a description of the spectacular high-pulsed emissions from oxide cathodes which were initially responsible for the success of magnetrons. Other matters which will be touched upon are decay of emission with time, cathode and anode sparking, the role of back-bombardment and secondary emission, and thoria cathodes.

ELECTRONIC FREQUENCY STABILIZATION OF MICROWAVE OSCILLATORS

R. V. Pound
(Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.)

For many purposes, microwave signal generators of very-high-frequency stability are desirable. Two methods of electronic frequency control relative to the resonant frequency of an isolated microwave cavity resonator are discussed. The relative frequency can be maintained to better than one part in 10^8 and the absolute frequency stability is that of the cavity. High-fidelity frequency modulation with deviations of the order of $f/14Q$ about the stabilized frequency is possible. Several uses are mentioned.

FROM WIRING TO PLUMBING

E. M. Purcell
(Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.)

At microwave frequencies circuit components are needed which, although comparable to a wavelength in size, provide the functional flexibility of ordinary radio-frequency circuit elements. Connectors, rotating joints, T-junctions, transformers, filters, variable reactors, attenuators, etc., must be manufacturable, durable, and in electrical behavior accurately predictable. In some respects microwave circuits are simpler than their low-frequency counterparts. Line-length transformations are conveniently and freely employed. Also, the symmetry properties of wave-guide transmission modes permit the realization, in simple form, of networks with interesting and useful properties. With a few notable exceptions microwave circuit design has been reduced to an engineering science.

TELEVISION-STUDIO EQUIPMENT

James J. Reeves
(Columbia Broadcasting System, New York, N. Y.)

The design of camera and studio control-room equipment, recently built and used by the Columbia Broadcasting System for the transmission of high-definition color television and 1029-line black-and-white images is described in detail.

MICROWAVE PROPAGATION

Part I

The Effect of Rain Upon the Propagation of Waves in the One- and Three-Centimeter Region

S. D. Robertson and A. P. King
(Bell Telephone Laboratories, Inc., New York, N. Y.)

The effects of rainfall and atmospheric absorption on the propagation of microwaves and the methods employed for measuring them are described. The attenuation of 3.2-centimeter waves is slight for moderate and light rainfall. During a cloudburst, the attenuation may approach a value of 5 decibels per mile. At 1.09 centimeters, the waves are appreciably attenuated even by a moderate rain.

Part II

Propagation of Six-Millimeter Waves

G. E. Mueller
(Bell Telephone Laboratories, Inc., New York, N. Y.)

Attenuations in excess of 25 decibels per mile have been observed during a cloudburst. The losses are still higher at 0.62 centimeter reaching a value of 42 decibels per mile for a cloudburst. The gas attenuation at this wavelength is probably less than 0.2 decibel per mile.

THE IMAGE ORTHICON, A SENSITIVE TELEVISION PICKUP TUBE

Albert Rose, P. K. Weimer, and H. B. Law
(RCA Laboratories, Princeton, N. J.)

The image orthicon is a new television pickup tube of high sensitivity. It operates stably at all light levels and the picture transmitted is substantially independent of the illumination over a wide range. Scenes have been transmitted successfully with one hundredth of the illumination previously required by commercial pickup tubes.

The advantageous features of the image orthicon have been attained by incorporating with the orthicon an electron multiplier, an electron-image section and a two-sided target. This new development is described in detail.

DUPLEX OPERATION OF INDEPENDENT HIGH-POWER OSCILLATORS FOR INDUCTION HEATING

W. C. Rudd
(Induction Heating Corporation, New York, N. Y.)

In the production of high-frequency power for induction heating, it has been found advantageous to use

multiple oscillators for the production of powers beyond that of a single unit. The problem involves the interconnection of two independent tank circuits in a manner such that the output frequency of the two units is the same and in synchronism. In addition, the total loan must be shared equally by the two power supplies. The paper describes the electrical problems involved, together with their practical solution for the production of powers of 40 and 50 kilowatts, using two 20- or two 25-kilowatt oscillators.

DESIGN CONSIDERATIONS IN BROADSIDE ARRAYS

John Ruze
(Evans Signal Laboratory, Belmar, N. J.)

The various factors considered in the design of broadside arrays are discussed and their effect on antenna bandwidth as determined by pattern deterioration are mentioned. Mutual impedance between antenna elements and means of establishing a desired current and phase distribution are detailed. Various methods of lobing a broadside array with magnitude of secondary lobes and bandwidth are discussed. Several designs will be detailed, such as the antenna system used on the SCR-270 and the AN/CPX-1.

A FREQUENCY-MODULATION ALTIMETER FOR METER AND LIGHT INDICATION AND THE AUTOMATIC ALTITUDE CONTROL OF AIRCRAFT

R. C. Sanders, Jr., and W. R. Mercer
(Raytheon Manufacturing Company, Watham, Mass.)
and
Irving Wolff and J. C. Smith
(RCA Laboratories, Princeton, N. J.)

AN AUTOMATIC VISUAL-INDICATING RADIO DIRECTION FINDER

Aldo Scandurra and Samuel Stiber
(Signal Corps Engineering Laboratory, Belmar, N. J.)

Radio set AN/TRD-2 is a Signal Corps development designed to provide a portable and mobile direction finder giving instantaneous indications and automatic sense. This direction finder is unique in that the instantaneous indication on the oscilloscope is a single ine generated from the center of the screen with a length proportional to the signal strength and the angular position of which indicates the direction of arrival of the signal without sense ambiguity. An additional feature is that noise and modulation are almost completely eliminated from the visual indication.

ELECTROOPTICAL CHARACTERISTICS OF TELEVISION SYSTEMS

O. H. Schade
(RCA Victor Division, Harrison, N. J.)

Optical and psychological capabilities of the human eye determine the performance standards for television systems.

Significant values for image detail and contrast, the electrical channel width, and signal-to-noise ratios are derived from the threshold visibility of picture detail and random brightness fluctuations.

Optical requirements and processes of developing electrical signals in television cameras for monochrome or color transmission are examined to establish relations and comparative values for obtainable signal-to-noise ratios and the required light flux.

SIGHT AND SOUND ON ONE CARRIER

Kurt Schlesinger
(Columbia Broadcasting System, New York, N. Y.)

The fundamental aspects of audio and video multiplex operation are outlined, and a description of the various forms of multiplex modulation is given. Theoretical and experimental considerations of performance characteristics are considered. There is a detailed discussion of transmitter and receiver circuits and a discussion of results of recent tests.

EQUIVALENT CIRCUITS FOR WAVE-GUIDE STRUCTURES

Julian Schwinger
(Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.)

Wave-guide structures can be completely and rigorously described by equivalent lumped-constant circuits. It is the purpose of the lecture to show how typical equivalent circuits are constructed by employing the special symmetry of the situation, to demonstrate that much qualitative information can be obtained from the form of the circuit even without knowledge of the values of the circuit constants, to discuss how the values of the circuit constants can be directly measured, and to indicate the procedure of the theory in calculating the values of the circuit parameters in their complete dependance on frequency and geometry.

DISCRIMINATORS FOR FREQUENCY-MODULATION RECEIVERS

S. W. Seeley
(RCA Laboratories, New York, N. Y.)

Many of the aspects of frequency-modulation detection, including the new amplitude insensitive discriminators, are discussed.

ULTRA-HIGH-FREQUENCY TELEVISION TRANSMITTERS AND ANTENNAS

Robert Serrell
(Columbia Broadcasting System, New York, N. Y.)

Ultra-high-frequency television transmitters of several types have been developed and built in the Columbia Broadcasting System laboratories. These transmitters embodying tubes developed during the war, utilize a 10-megacycle modulation bandwidth. The design and performance of these transmitters and of wide-band antennas used will be described in detail. Also, some results of experimental ultra-high-frequency broadcasts are discussed.

MEASUREMENT OF THE ANGLE OF ARRIVAL OF MICROWAVES

W. M. Sharpless
(Bell Telephone Laboratories, Inc., New York, N. Y.)

This paper describes a method of measuring the direction from which microwaves arrive at a given receiving site. Data which have been collected on two short optical paths using a wavelength of $3\frac{1}{4}$ centimeters are presented to illustrate the use of the method. Angles of arrival as large as $\frac{1}{2}$ degree above the true angle of elevation have been observed in the vertical plane while no variations greater than $\pm 1/10$ degree have been found in the horizontal plane. More recent work, using a lens-type scanning antenna operated at a wavelength of $1\frac{1}{4}$ centimeters revealed that, at times, as many as four distinct transmission paths were present simultaneously on a 12.6-mile circuit. Simultaneous meteorological soundings were made near both terminals of the circuit.

MODEL AIRCRAFT-ANTENNA MEASUREMENTS

George Sinclair, E. W. Vaughan, and Edward C. Jordan
(Ohio State University Foundation, Columbus, Ohio)

Although antenna models have been used for many years in studying antenna patterns, the measurements were severely limited as to frequency and type of antenna. It is shown how the techniques may be extended to cover a very wide frequency range and a wide variety of antennas. Particular application to the study of aircraft antenna patterns are cited. The utility of models in studying special properties of antennas, such as polarization errors of direction finders, propeller-modulation effects, ellipticity of polarization, etc., are discussed.

ONE-MILLIONTH-OF-A-SECOND RADIOGRAPHY AND ITS APPLICATIONS

C. M. Slack and D. C. Dickson
(Westinghouse Electric Corporation, Bloomfield, N. J.)

The making of ultra-speed radiographs, using exposure times of the order of one millionth of a second, requires the passage of electron currents of 1000–2000 amperes. Such currents can be supplied by an electron source utilizing field emission from a cold-cathode electrode which degenerates into a metallic arc in a high vacuum. The recording of such high-speed transients are briefly discussed. The development of this equipment has been greatly accelerated because of the war. Slides showing its applications to various radiographic problems requiring short exposure times which have just been released by the War Department are shown; among these are radiographs taken at Frankford Arsenal and Aberdeen Proving Grounds of exploding shells and bombs, and at Princeton University, showing the wounding mechanism of high-velocity fragments. Future applications also are discussed.

THEORY OF IMPULSE NOISE IN IDEAL FREQUENCY-MODULATION RECEIVERS

David B. Smith and W. E. Bradley
(Philco Corporation, Philadelphia, Pa.)

An analytical treatment of noise in frequency-modulation receivers is given for the case when a mixture of noise and useful signals is applied to the receiver input. Methods of measuring the performance of receivers with respect to frequency-modulation noise are discussed.

GROUND-CONTROLLED APPROACH

Ernest Storrs, W. Devitt, and Ben Green
(Watson Laboratories, Red Bank, N. J.)

Out of the development race for military supremacy in World War II has come one of the most outstanding navigational aids yet known. It is known as ground-controlled approach or, more popularly termed, GCA. The military designation has been the nomenclature AN/MPN-1. Its purpose is to talk the pilot down the glide path under conditions approaching zero visibility to a safe landing by means of the normal communications equipment already installed in the aircraft.

A KINESCOPE FOR HOME PROJECTION-TYPE TELEVISION RECEIVERS

L. E. Swedlund
(RCA Victor Division, Harrison, N. J.)

The development of a small high-voltage cathode-ray tube for home projection-type television receivers,

soon to be available commercially, is described. Several new insulation and fluorescent screen problems had to be solved because of the need for operating at relatively high voltage. An outstanding gain in light output and performance was attained by applying an electron-transparent, light-reflecting aluminum film to the back of the fluorescent screen.

HIGH-FREQUENCY PLATED QUARTZ-CRYSTAL UNITS FOR CONTROL OF COMMUNICATIONS EQUIPMENT

R. A. Sykes
(Bell Telephone Laboratories, Inc., New York, N. Y.)

A description is given of the general problems relating to the development of high-frequency plated-crystal units, the methods employed for supporting the crystal blank, adjustment to final frequency by means of vaporized or evaporated films, and the various problems associated with the aging of such units. The general problems of devising simplified procedures for the large-scale manufacture of such crystal units are also described.

CRYSTAL RECTIFIERS IN HETERODYNE RECEIVERS

H. C. Torrey
(Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.)

Crystal rectifiers as frequency converters are superior to other detectors of microwave signals as measured by the noise figure—a quantity dependent on conversion loss and noise temperature of the mixer. Linear-network theory can be applied to microwave converters, a mixer being represented as a three-terminal pair device with terminals at signal, image sideband, and intermediate frequency. Such a representation is helpful in estimating the effect on conversion loss of image-sideband termination and of parasitic impedances. Other factors of importance are resistance to burnout and stability. Representative microwave mixers designed by Radiation Laboratory are described.

RADAR ASPECTS OF NAVAL FIRE CONTROL

Captain Dundas P. Tucker
(U. S. Navy, Bureau of Ordnance, Navy Department, Washington, D. C.)

Application of radar to fire control required the development of many new designs and techniques for the purpose of providing high precision in measuring target position and high resolution against multiple targets and interfering objects. A special circuit for measuring time intervals to 0.05 microsecond, and angles to 0.05 degree were developed. American possession of radar-controlled gunfire was a decisive factor in many naval engagements with Japanese ships and aircraft.

BEAM-SHAPING METHODS IN ANTENNA DESIGN

L. C. Van Atta
(Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.)

The angular width of an optical searchlight beam is determined by the extended light source placed at the focal point of the paraboloid reflector and by inaccuracies in the reflector shape. The beam width of a microwave paraboloid antenna, however, is due to diffraction of the radiation at the aperture. For sharp beams, it is desirable, frequently, to distribute the radiation in some manner other than that determined by diffraction. The focusing property of the antenna system can be modified either by providing an extended antenna feed or distorting the paraboloid shape. Straightforward experimental and theoretical approaches are available to the antenna designer in achieving a wide variety of beam shapes for special application.

STAGGER-TUNED WIDE-BAND AMPLIFIERS

H. Wallman
(Radiation Laboratory, Massachusetts Institute of Technology, Cambridge, Mass.)

In radar and television receivers, the simplicity of single-tuned circuits strongly commends itself to the designer of wide-band, band-pass amplifiers. Unfortunately, with present tubes, one cannot make a high-gain amplifier wider than about 4 megacycles; the principal reason is the rapidity with which bandwidth decreases as identical single-tuned stages are cascaded. This paper describes the scheme of stagger-tuning the individual single-tuned circuits, which essentially eliminates the shrinking of over-all bandwidth and thus makes possible very simple 10- and 15-megacycle wide amplifiers. Stagger-tuning appears to be much less well known than it deserves. Graphs are presented condensing the whole basic design of a wideband amplifier into the work of a few minutes; actual examples of amplifiers are shown.

METALLIZED-GLASS ATTENUATORS FOR RADIO-FREQUENCY APPLICATIONS

Ernst Weber
(Polytechnic Institute of Brooklyn, Brooklyn, N. Y.)

Based on special techniques of precision metallizing of glass plates and tubes, power-absorbing elements have been developed suitable for use in calibrated attenuators in coaxial as well as wave-guide radio-

frequency power transmission systems of moderate power level. Particular attention had to be given to proper matching of the attenuating sections to the practically lossless transmission lines by means of impedance transformers which themselves are part of the metallized-glass elements. Several models with precision measuring drives were developed for use as reference standards.

THE NEW "SPEEDOMAX" POWER-LEVEL RECORDER

A. J. Williams, Jr., and W. R. Clark
(Leeds and Northrup Company, Philadelphia, Pa.)

The new "Speedomax" power-level recorder is an instrument that faithfully records rapid or slow variations in power level with time, and is independent of variations in frequency within its frequency range (25 to 150,000 cycles). The recorder scale is linear in decibels. A 20-decibel change in power input can be recorded in about one second. The new circuits used in this new recorder are described in detail, and its performance characteristics are given.

A PULSE ALTIMETER OF HIGH ACCURACY AT HIGH ALTITUDES

Irving Wolff, W. D. Hershberger, G. W. Leck and R. R. Welsh
(RCA Laboratories, Princeton, N. J.)

COLOR-TELEVISION TRANSMITTER

N. H. Young
(Federal Telecommunication Laboratories, Inc., New York, N. Y.)

On December 21, 1945, Federal Telecommunication Laboratories delivered to the Columbia Broadcasting System a color-television transmitter. It operates on a carrier frequency of 490 megacycles with a peak power output of 1 kilowatt and can be modulated uniformly at all frequencies from direct current to 10 megacycles. It is the most powerful transmitter of this frequency and bandwidth ever used in television service.

The transmitter is made possible by the development of a new ultra-high-frequency triode, type 6C22. This tube is used in the four final stages of the radio-frequency section and in the two final stages of the video-frequency section.

A conventional sequence of crystal oscillators and frequency multipliers is used in the radio-frequency section. The video-frequency section is unusual chiefly in the use of direct-current coupling between amplifier stages, thus making the use of direct-current restoring circuits unnecessary. The final radio-frequency amplifier is grid-modulated, and the low plate resistance of the tube aids greatly in securing adequate bandwidth in the output radio-frequency circuit.

HERMETICALLY SEALED METAL HOLDER FOR CRYSTAL UNITS

A. W. Zeigler
(Bell Telephone Laboratories, Murray Hill, N. J.)

A metal-type holder has been developed to replace the phenolic-type for hermetically sealing low- and high-frequency crystal units for use by the various branches of the Armed Forces. The holder comprises an assembly of a drawn-metal cover fluxless soldered by induction heating to a base carrying glass-to-base seal terminals. The techniques developed follow closely those used in miniature vacuum-tube assemblies. Similar holders are being developed for peacetime applications in the communication field.

Awards

W. C. White, chairman of the Awards Committee, presented to the Board of Directors on November 7, 1945, the list of nominees selected for the 1946 Awards. These recommendations were unanimously approved by theBoard, and the Awards were presented at the Banquet of the Winter Technical Meeting held on Thursday, January 24, 1946.

Medal of Honor—1946

Ralph Vinton Lyon Hartley

"For his early work on oscillating circuits employing triode tubes and likewise for his early recognition and clear exposition of the fundamental relationships between the total amount of information which may be transmitted over a transmission system of limited bandwidth and the time required."

Fellow Award—1945

Thomas Lydwell Eckersley

"For his outstanding contributions to the theory and practice of radio-wave-propagation research. Both his approach to the problem from the standpoint of practical communications and his invention of mathematical tools useful in the computation of radiated fields are achievements of lasting value acclaimed by the whole radio world, and form a monument of which he may be justly proud."

Fellow Award—1945

Gregory Breit

"For pioneering in the experimental probing of the ionosphere and giving to the world the first publication of the experimental proof of the existence of the ionosphere; and for having initiated at an early date the pulse method of probing by reflection which is the basis of modern radar."

Morris Liebmann Memorial Prize—1945

Peter C. Goldmark

"For his contributions to the development of television systems, particularly in the field of color."

Fellow Award—1945

Harold Lester Kirke

"For his services to broadcasting in the British Isles and in particular for his leadership in the research activities of the British Brcadcasting Corporation."

Fellow Award—1945

Joseph Slepian

"In recognition of his contributions as scientist, engineer, and inventor to the field of electronics."

Fellow Award—1945

Clarence W. Hansell

"For his pioneer work in the development and application of equipment for the ever higher frequencies employed for radio communication."

Fellow Award—1945

Henri G. Busignies

"For his accomplishments in the field of radio direction finders, particularly pioneering work on instruments having automatic indicating features."

Conway Studios, Inc.

Fellow Award—1945

Harold S. Osborne

"For his contributions in the electrical communication field including outstanding leadership and direction in the application of new techniques to telephony."

Fellow Award—1945

HOWARD A. CHINN

"For his contributions to improved broadcasting."

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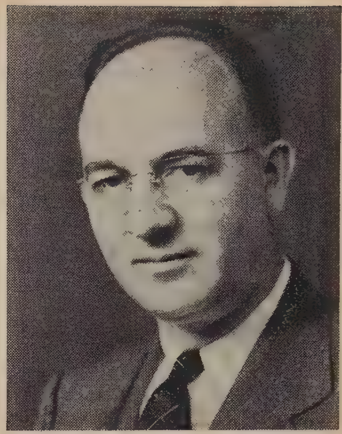
Fellow Award—1945

WALTER C. EVANS

"In recognition of his past contributions to radio and his present active participation in the affairs of the Institute."

❖



Fellow Award—1945

ELMER D. MCARTHUR

"For his developments in the field of ultra-high-frequency electron tubes."

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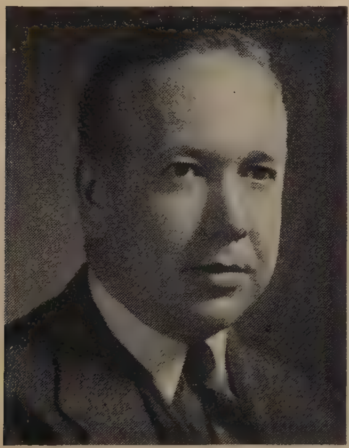


Fellow Award—1945

JULIUS A. STRATTON

❖

JULIUS A. STRATTON

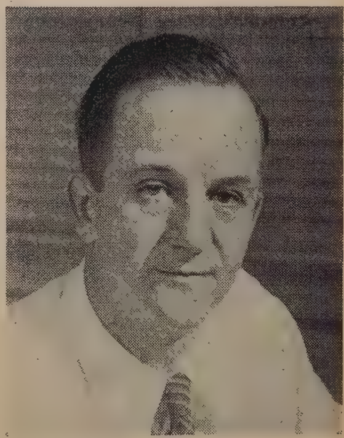
"In recognition of his contributions as a teacher and author, adapt in the field of fundamental research who has applied his knowledge to improve radio communications."

❖

MERLE ANTHONY TUVE

"For pioneering in the experimental probing of the ionosphere and giving to the world the first publication of the experimental proof of the existence of the ionosphere; and for having initiated at an early date the pulse method of probing by reflection which is the basis of modern radar."

❖



Fellow Award—1945

MERLE ANTHONY TUVE

❖

Fellow Award—1945

ARTHUR L. SAMUEL

"For his fundamental work in the field of electronic research and for development of electronic devices of particular value at very high frequencies."

Fellow Award—1945

WILLIAM O. SWINYARD

"In recognition of his work in promoting electronics and the affairs of the Institute, particularly in his district."

K. Rarich

Fellow Award—1945

RONALD J. ROCKWELL

"For active work in the affairs of the Institute and in the engineering of high-power international broadcast transmitters."

Institute News and Radio Notes

Board of Directors

December 5 Meeting: At the regular meeting of the Board of Directors, which was held on December 5, 1945, the following were present: W. L. Everitt, president; G. W. Bailey, executive secretary; S. L. Bailey, W. L. Barrow, L. M. Clement, W. H. Crew, assistant secretary; R. F. Guy R. A. Hackbusch, R. A. Heising, treasurer; Keith Henney, F. B. Llewellyn, B. E. Shackelford, D. B. Sinclair, H. M. Turner, H. A. Wheeler, and W. C. White.

Executive Committee Actions: It was unanimously approved that the actions of the Executive Committee taken at its November 7, 1945, meeting, be ratified.

Committees

Building-Fund: Dr. Shackelford, chairman of the I.R.E. Building-Fund Committee, reported that the Fund had passed $623,000, and that the final report of the Building-Fund Committee would be submitted at the January meeting.

Tellers: The report of the Tellers Committee, dated November 15, 1945, was unanimously approved.

Constitution and Bylaws

Executive Committee: It was unanimously approved to adopt, as of December 3, 1945, the following amended Bylaw, Sec. 27:

"Sec. 27—The Executive Committee shall be responsible for the management of the office of the Institute, which shall include special activities, technical activities, office operations, accounting, advertising, and publications, and the setting of salaries (with the exception of Officers' salaries) in all instances in the limits of the budget. Such activities as the Executive Committee shall direct shall be administered by Executive Secretary. The salary of the Executive Secretary shall be determined by the Executive Committee."

Price of PROCEEDINGS *in Annual Dues:* It was unanimously approved to adopt, as of December 3, 1945, the following amended Bylaw Sec. 57:

"Sec. 57—The price of a single annual subscription to the PROCEEDINGS for a Fellow, Senior Member, Member, or Associate shall be $6.00, to be included in his annual dues as specified in Article IV of the Constitution."

Regional Representation Plan: The Board considered the proposed Constitutional Amendments for the Regional Representation Plan and made suggestions for minor changes which will be made in the final draft by the Constitution and Laws Committee. The following actions were taken:

a. The term "Regional Committee" was adopted.

b. The entire group of proposed Constitutional Amendments for Regional Representation will be submitted to the membership as the only option, and modifications in the Constitution which are desirable, but not essential, to the Plan may also be included in the group.

Student Status: It was unanimously approved that the Board rescind the previous motion of October 3, 1945, approving two and one-half years in Student status.

Postponement of Dues: It was unanimously approved that the Board approve the recommendation of the Executive Committee that the final date of payment of 1946 dues outside of United States and Canada be extended from April 30 to June 30, 1946.

Appointment: Dr. Alfred N. Goldsmith was reappointed as I.R.E. representative on the Standards Council, and Mr. Raymond F. Guy as alternate, for the three-year period, 1946–1948.

Executive Committee

December 5 Meeting: The Executive Committee meeting, held on December 5, 1945, was attended by W. L. Everitt, president; G. W. Bailey, executive secretary; S. L. Bailey, W. L. Barrow, E. F. Carter, W. H. Crew, assistant secretary; R. A. Heising, treasurer; and F. B. Llewellyn (guest).

Membership: Approval was given to the 390 applications for membership in the Institute listed on page 46 A of the January, 1946, issue of the PROCEEDINGS of the I.R.E. and WAVES and ELECTRONS. These applications are as follows:

For Transfer to Senior Member grade	15
For Admission to Senior Member grade	7
For Transfer to Member grade	52
For Admission to Member grade	53
For Admission to Associate grade	169
For Admission to Student grade	94
Total	390

Committees

The following appointments were approved:

EDUCATION

R. W. Jones, Elmer H. Schulz, Glenn Koehler, Robert P. Siskind, Jack D. Ryder, Karl Spangenberg, Alexander H. Wing, Jr.

FACSIMILE

E. F. Watson

PIEZOELECTRIC CRYSTALS

Paul L. Smith

TELEVISION

W. F. Bailey, S. Maulner, M. E. Strieby

RADIO RECEIVERS

J. M. Pettit

RADIO WAVE PROPAGATION

Earl Cullum

URSI: It was unanimously approved that the Institute of Radio Engineers cooperate with the International Radio and Scientific Union in the Spring Meeting.

Sixth Annual Broadcast Engineering Conference

Under the joint sponsorship of The Ohio State University and the University of Illinois and with the continued co-operation of the National Association of Broadcasters and The Institute of Radio Engineers, the Broadcast Engineering Conference will be resumed this year. This is a continuation of the annual conferences held during the years from 1938–1942, inclusive. Dr. W. L. Everitt, now head of the department of electrical engineering at the University of Illinois, Urbana, Illinois, will continue to act as the director with Professor E. M. Boone of The Ohio State University as associate director.

The 1946 conference will beheld at The Ohio State University in Columbus, Ohio, during the week of March 18–23. The conference will be held annually, and the place of meeting will alternate between the campus of The Ohio State University and that of the University of Illinois. Emphasis in the program will be placed on the impact of developments since 1942 on operating problems in broadcast engineering, including frequency modulation and television.

As a result of the war, many engineers have moved or changed their affiliations, and many new men have entered the field. The mailing list accumulated during previous conferences is obviously out of date. The director requests that those interested notify him of their present address so that they can be informed of the details of the program as soon as available. Communications to Dr. Everitt should be addressed c/o University of Illinois, Urbana, Illinois.

PROGRAM

All meetings to be held in Campbell Hall Auditorium, The Ohio State University

Monday, March 18

9:00 A.M. to 1:00 P.M.

"Contributions of War Developments to Broadcasting" by A. B. Chamberlain, Chief Engineer, Columbia Broadcasting System

"Symposium on Broadcast Maintenance Problems," by A. J. Ebel, Chief Engineer, University of Illinois Radio Service, *Chairman*

2:30 P.M. to 4:30 P.M.

"Design of Broadcast Studios with Irregular Boundary Surfaces," by K. C. Morrical and J. E. Volkman, Radio Corporation of America

Tuesday, March 19

9:00 A.M. to 1:00 P.M.

"Antenna Patterns and the Antennalyzer," by George H. Brown, Research Engineer, Radio Corporation of America

"Symposium on Recording Techniques," by Lynn Smeby, Associate Director, Operational Research Staff, Office of the Chief Signal Officer, U. S. War Department

2:30 P.M. to 4:30 P.M.

General Acoustical Problems in Broadcasting," by E. J. Content, Station WOR

Wednesday, March 20

9:00 A.M. to 1:00 P.M.

"Symposium on Very-High-Frequency Antenna and Coupling Circuits," by E. C. Jordan, Department of Electrical Engineering, University of Illinois, *Chairman*

"Symposium on Television-Station Operation," by Robert E. Shelby, National Broadcasting Company, *Chairman*

2:30 P.M. to 4:30 P.M.

"Radio Relays for Frequency Modulation and Television"

Thursday, March 21

9:00 A.M. to 1:00 P.M.

"Stratovision," by Ralph Harmon, Westinghouse Electric Corporation, and representatives from Glenn L. Martin Aircraft Co.

"Round Table and Question Box," by A. D. Ring, Consulting Engineer, *Chairman;* John Willoughby, Assistant Chief Engineer, Federal Communications Commission, in charge of Broadcasting; also representative chief engineers from broadcasting stations.

2:30 P.M. to 4:30 P.M.

"Interconnecting Facilities for Frequency Modulation and Television Broadcasting," by H. I. Romnes and W. E. Bloecker, American Telephone and Telegraph Company

Friday, March 22

9:00 A.M. to 1:00 P.M.

"High-Powered Tubes for Very-High-Frequency Operation," by W. W. Salisbury, Collins Radio Company

"Symposium on Frequency-Modulation Operating Problems," by Philip B. Laeser, *Milwaukee Journal* Company, *Chairman*

2:30 P.M. to 4:30 P.M.

"Symposium on Frequency-Modulation Monitors," by R. C. Higgy, Director WOSU, Ohio State University, *Chairman;* D. B. Sinclair, General Radio Company; Frank Gunther, Radio Engineering Laboratories; H. R. Summerhayes, Jr., General Electric Company

Saturday, March 23

9:00 A.M. to 1:00 P.M.

"Symposium on Frequency-Modulation Methods," by W. L. Everitt, Head, Department of Electrical Engineering, University of Illinois, *Chairman*

"Symposium on Field Experiences in Very-High-Frequency Propagation," by Raymond M. Wilmotte, Consulting Engineer, *Chairman*

Tuesday, March 19

8:00 P.M.

Popular Scientific Lecture, University Hall, The Ohio State University

Thursday, March 21

6:30 P.M.

Banquet, Fort Hayes Hotel

Radio Pioneers' Dinner

The New York Section of the Institute of Radio Engineers held its fourth Radio Pioneers' Dinner at the Hotel Commodore in New York City on the evening of November 8, 1945. The General Committee, under the chairmanship of Louis Gerard Pacent, planned ambitiously and effectively; the attendance of over 800 eclipsed by far any other dinner meeting of the Section and has been exceeded only seldom at national convention banquets of the Institute.

The American Radio Relay League, Radio Club of America, and Veteran Wireless Operators Association were guests of the Section and were represented by their presidents: George Bailey, also executive secretary of the Institute; Fred Klingenschmitt; and William J. McGonigle, respectively. President Everitt of the Institute was master of ceremonies.

Although the keynote of the evening was fun, John V. L. Hogan, who together with Robert H. Marriott and Alfred N. Goldsmith was responsible for the merging of the Wireless Institute and the Society of Wireless Telegraph Engineers into the Institute, read excerpts of a brief address he prepared for the occasion. His theme, "radio will be developed faster if those engaged in it will work together more," was taken from a letter sent by Mr. Marriott in 1909 to some 200 wireless people who, it was hoped, would be interested in founding a radio society, the Wireless Institute.

Mr. Marriott, first president of both the Wireless Institute and the Institute of Radio Engineers, and Dr. Goldsmith, indefatigable worker and supporter of the Institute and editor of its publications, were unable to be present. Their messages, read by President Everitt, did much to soften the reality of their absence.

Held during the week commemorating the twenty-fifth anniversary of the initiation of radio broadcasting, it was fitting for The American Broadcasting Company to arrange for Deric Leighton to direct a skit depicting some of the early trials of "Wireless Pioneering," for Ed East to entertain at the piano, and for Ray Knight to tell of the "Early Days and Late Nights."

Some twenty gifts were contributed as door prizes. Included among them was an unannounced surprise which was won by General Van Deusen—a week-old bull calf—pointfree! As the prizes were drawn, they were augmented by others put up by those present. Among the "upsets" were the winning of a General Electric receiver by Louis Pacent, Jr., presented by the chairman of the Prize Committee, his boss, Dorman Israel, vice-president of Emerson, and the selection of E. M. Deloraine, vice-president-elect of the Institute and president of the International Telephone and Telegraph International Telecommunication Laboratories, as the recipient of a National communications receiver.

Sergeant Irving Strobing, who transmitted the last message from Bataan before it was overrun by the Japanese, was introduced and expressed his desire to continue his radio studies. Jack Poppele, chairman of the Scholarship Committee of the Veteran Wireless Operators Association, presented him with a scholarship in the Capital Radio Engineering Institute, an on-the-spot decision of the officers of that organization.

Reproduced on the editorial page of this issue, are messages from two of the honored guests who were present, Admiral Redman and General Ingles, heads of communications of the Navy and Army. The significance and scope of their tributes to all radio-and-electronic engineers for outstanding accomplishments in the war efforts, place their statements far beyond the confines of any "old-timers dinner."

Jack Binns, famed radio operator of the

Louis G. Pacent, general chairman of the Radio Pioneers' Dinner of the New York Section presenting the gavel of master of ceremonies to Dr. Everitt, president of the Institute.

The group responsible for the Radio Pioneers' Dinner is shown above. Around the table from left to right are George B. Hoadley, chairman of the New York Section; Louis G. Pacent, general chairman of the Dinner Committee; W. L. Everitt, president of the Institute and master of ceremonies; Roger M. Wise; Glen Payne; Raymond Guy; Ralph R. Batcher, secretary; Edward J. Content, treasurer; D. W. May; P. R. Morton; J. C. Stroebel; H. C. Gawler; and E. L. Bragdon. Missing were vice-chairman George Lewis, O. H. Caldwell, George H. Clark, George C. Connor, John Di Blasi, Hugo Gernsback, O. B. Hanson, J. Q. A. Holloway, Robert H. Marriott, Dorman D. Israel, Harry Sadenwater, and Harold P. Westman.

S.S. Republic disaster, tried his fist, not too successfully, on the first radio set offered for sale to the general public. Built by the Electro Importing Company in 1909, its spark-coil transmitter and coherer-and-electric-bell detector were too slow for Binns' still-modern fist. The presence of various representatives of the Federal Communications Commission was evidence of the wisdom of Hugo Gernsback in obtaining a special license for the operation of the transmitter at the dinner.

A souvenir booklet, unique perhaps in that it carried no advertising, was devoted to a seriously presented history of wireless from its early beginnings through 1925. Designed to stimulate the memories of all old-timers, it contained numerous pictures of equipment and personalities inextricably associated with the "good old days."

Miniature replicas of volume I, number 1, of *Modern Electrics*, published in April, 1908, were also distributed through the courtesy of Hugo Gernsback. This was the first regular magazine in America devoted to wireless.

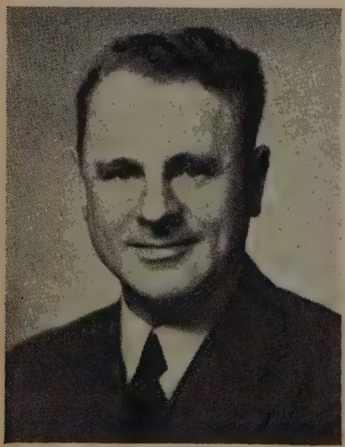
Twenty-seven commercial organizations made financial contributions to the success of the meeting. The General Committee planned so well that not only was the deficit from the cocktail party, dinner, and souvenir booklet covered, but the New York Section was enabled to make a substantial contribution to the Building Fund of the Institute.

I.R.E. People

Canadian Short-Wave Broadcasting

A new short-wave international broadcast station has been established by the Canadian Broadcasting Corporation near Sackville, New Brunswick. The equipment includes two 50-kilowatt short-wave transmitters. Space is available for an additional 7.5-kilowatt short-wave transmitter.

Installation of the equipment was supervised by Joseph M. Conroy (A'23) and Fred R. Quance (A'42), engineers of the RCA Victor Division at Camden, New Jersey. The layout of the building housing the equipment was carried out by a group of engineers under the direction of G. W. Olive (A'29) of the same organization. Numerous technical problems of interest were encountered and solved in the design and installation of this new Canadian Station CHTA.



G. W. Olive

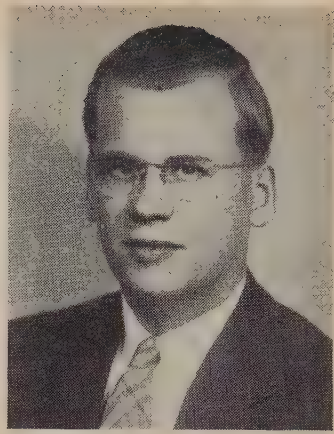
Fred R. Quance

Joseph M. Conroy

Harris and Ewing

Raymond M. Wilmotte

Paul H. Crago



G. Curtis Engel

New Wilmotte Laboratory

Raymond M. Wilmotte (M'29-F'38), director of Wilmotte Laboratory, Inc., of Washington, D. C., recently announced the opening of a branch laboratory at 236 West 55 Street, New York 19, N. Y., offering engineering services in the application of electronics to industrial uses.

During the war, Wilmotte Laboratory devoted its activities largely to research and development for the Army, Navy, and Office of Scientific Research and Development. Much of this work was on radar, fire control, and other special devices. In the course of this work, the Wilmotte Manufacturing Company was organized to produce special electronic devices.

It is Mr. Wilmotte's intention to expand this combination of facilities in order to provide industrialists with complete engineering services applying the full potentialities of electronics to their processes.

The New York laboratory will be under the supervision of Paul H. Crago (A'44), who was section engineer with the Union Switch and Signal Company, in charge of design and application of electromagnetic signaling devices for railroads. He was more recently with Specialties, Inc., as executive engineer in charge of research and development of ordnance instruments for the Armed Forces, and the Federal Telephone and Radio Corporation.

G. Curtis Engel (A'45), formerly production manager of General Electronics, Inc., and manager of the laboratory of General Time Instruments Corporation, will head a group specializing in electronic controls and inspection methods. Another group will do consulting work in the application of dielectric heating.

E. C. Page and J. Wesley Koch

Lieutenant Colonel E. C. Page (M'42–SM'43), director of engineering of the Mutual Broadcasting System, has announced that Captain J. Wesley Koch (J'29–S'33–A'36–M'45) joined the engineering staff of the network on October 15 when he was released from the Army.

While in the Army, Captain Koch served as technical adviser of the Radio Propagation Unit and was concerned with radio propagation conditions throughout the world. He was a staff radio officer at Allied Force Headquarters in North Africa and Italy, and until October, 1945, he was with the office of the Chief Signal Officer at Baltimore, Maryland.

Captain Koch is a native of Nebraska and received his B.S. degree in electrical engineering at the University of Nebraska in 1934. It was while he was at the University that he designed and constructed special equipment for a wired-radio program distributing system operated by the Program Service of Lincoln, Nebraska. He later accepted a position with KFEQ, St. Louis, Mo., in 1934, as chief engineer. It was here that he designed and constructed most of the transmitting equipment at this station. After installing a complete new transmitting plant at KFEQ in 1942, Captain Koch resigned to accept a commission with the Signal Corps.

E. C. Page

J. Wesley Koch

Jack Kaufman

Jack Kaufman (A'30–SM'44) has been appointed a vice-president of the Aireon Manufacturing Corporation. He recently joined the organization to head the new San Francisco office which serves the transportation and communications industries on the Pacific Coast.

A graduate of the University of California in 1917, Mr. Kaufman has been engaged for a number of years in electronics in San Francisco. He was formerly president of Heintz and Kaufman, Limited, and executive vice-president of Globe Wireless, Ltd.

Mr. Kaufman formerly was president of the West Coast Electronic Manufacturers' Association, San Francisco Council, and vice-president of the coast-wide group of the same Association. Until recently, he was a member of the Industry Advisory Committee with the Board of War Communications.

Fabian Bachrach

HUGH S. KNOWLES

Hugh S. Knowles

Hugh S. Knowles (A'25–F'41), vice-president and chief engineer of the Jensen Radio Manufacturing Company, recently was elected president of the Acoustical Society of America.

Before graduating from Columbia University, Mr. Knowles was an editor of *Popular Radio;* technical columnist for the *New York Herald Tribune;* manager of Columbia University's experimental radio station; and president of the University's Radio Club. After receiving his Bachelor's degree, he undertook further work at Columbia and extensive graduate work at the University of Chicago. As a guest lecturer at Chicago University, he prepared and gave a graduate course on sound and vibrating systems. Mr. Knowles then served as radio engineer for Hammarlund Manufacturing Company and as manager of the parts division of Silver-Marshall. In 1931, he joined the staff of the Jensen Radio Manufacturing Company as chief engineer and was elected to the position of vice-president in charge of product research and development in 1940.

Mr. Knowles was chairman of the Committee on National Defense, the Electroacoustics Standards Committee of The Institute of Radio Engineers, and Sound Equipment Standards Committee of the Radio Manufacturers Association. He has also represented the IRE and the RMA on various electroacoustical committees of the American Standards Association, and he served as chairman of the Chicago Section of the IRE. Mr. Knowles received the Fellowship Award of the IRE for "engineering leadership in the field of acoustics and its radio applications."

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Norman B. Neely

Norman B. Neely (A'39), West Coast manufacturers' representative for several electronic firms, has resumed active management of his firm, Norman B. Neely Enterprises of Hollywood, California. During the war, Mr. Neely was associated with Western Electric Company as special field engineer engaged in confidential work on Army Air Force equipment contracts.

Paul D. Zottu

Paul D. Zottu (A'31–M'38–SM'43–F'44), formerly chief engineer, Thermex Division, the Girdler Corporation, Louisville, Kentucky, has announced his entrance into the field of consulting industrial electronic engineering. He offers to industrial users and equipment manufacturers a consulting service in the field of high-frequency induction and dielectric heating, specializing in applications, equipment and component design, and equipment selection.

Mr. Zottu has an extensive background in the radio field covering a wide range of activities. The last seven years, outside of the part time spent at the Radiation Laboratory at Massachusetts Institute of Technology, have been devoted entirely to industrial high-frequency heating problems. In this field, he pioneered the design and application of many industrial installations. Mr. Zottu is chairman of the Technical Committee on Industrial Electronics of the I.R.E., a member of the committee on Industrial Heating Applications of the Radio Technical Planning Board, Induction and Dielectric Heating Committee of the American Institute of Electrical Engineers, chairman of the committee on Frequency Allocations of the Society of the Plastics Industry, member of the American Physical Society, American Association for the Advancement of Science, and Society of the Plastics Industry.

PAUL D. ZOTTU

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Allan R. Ogilvie

Allan R. Ogilvie (A'43) has been named a vice-president of Maguire Industries, Inc., in charge of its Bridgeport, Connecticut plant.

A native of Boston, Mr. Ogilvie studied at the Massachusetts Institute of Technology and Temple University. His first position was with the Wireless Specialty Company, after which he became affiliated with Western Electric Company. Later, Mr. Ogilvie joined the Radio Corporation of America where he was an engineer for fourteen years. He became a member of the staff of Maguire Industries, Inc., in 1944, and was chief engineer of the Company's electronics division.

D. L. JAFFE

D. L. Jaffe

D. L. Jaffe (A'37) has joined the Polarad Electronics Company of New York as a partner and general manager.

From 1937 to 1939, Dr. Jaffe performed research in frequency modulation at Columbia University, and in 1939, he joined the Columbia Broadcasting system as television engineer engaged in studio-equipment development. Dr. Jaffe was appointed development engineer in the microwave radar division of the Raytheon Manufacturing Corporation in 1942, during which time he acted as liaison engineer between that Corporation and the Radiation Laboratory at the Massachusetts Institute of Technology. He then joined the staff of the Templetone Radio Manufacturing Corporation as chief research engineer.

Dr. Jaffe is an Associate member of the American Institute of Electrical Engineers and a member of Sigma Xi.

❖

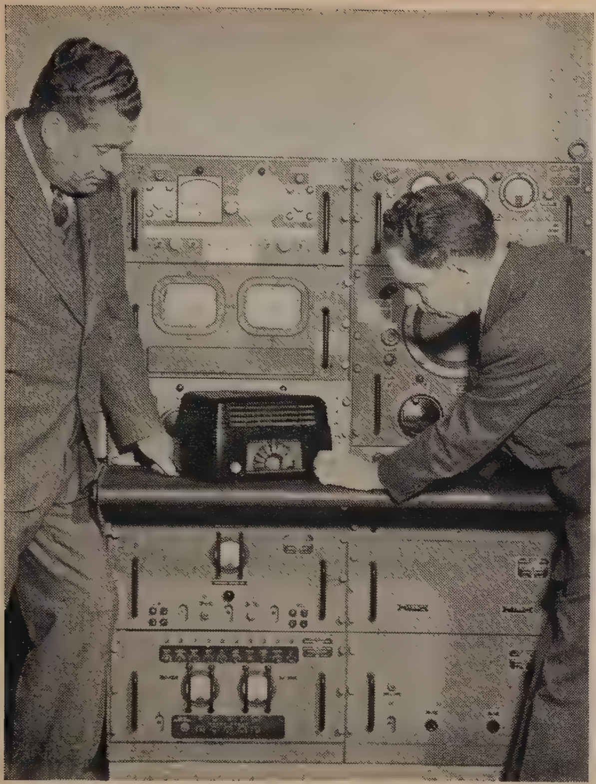
Claude Lyons

Claude Lyons (M'27-SM'43) joint managing director of Messrs. Claude Lyons, Ltd., Liverpool and London, England, was appointed on July 5, 1945, as the first Chairman of the Federation of Anglo-American Importers, the headquarters of which are at 180 Tottenham Court Road, London, W.1, England. The Federation is a nonprofit organization open to importers into England of radio, electronic, and allied equipment; manufacturers of such equipment in the United States of America; and persons interested in the promotion of and the continued welfare of the radio, electrical, and allied industries, as well as good relationships in international commerce.

❖

F. C. McMullen

F. C. McMullen (A'42) has been appointed chairman of the aviation section of the Radio Manuacturers Association's transmitter division, and is in charge of aviation radio sales for Western Electric Company. He succeeds J. W. Hammond (A'44–SM'45) of Bendix Radio, Baltimore, Maryland.



Peacetime Radio Production

I. J. Kaar (right) (J'22–A'24–M'29–F'41), manager of General Electric's receiver division, discusses first peacetime radio off production lines at the company's Bridgeport, Connecticut, plant, with Paul L. Chamberlain (A'43), manager of sales for the division. In background is radar console.

Henry Grossman

Progress in the various phases of television research of the Columbia Broadcasting System, under the direction of Peter C. Goldmark (A'36–M'38–F'42), has made possible an integration of technical operations in the television field with the network's other New York broadcast operations. Accordingly, Henry Grossman (M'34–SM'43) will share in the responsibility for television technical operation routines, personnel, and routine maintenance of equipment, supplementing his previous broadcast commitments. The responsibility for equipment design and installation, engineering standards, and development tests will remain under Dr. Goldmark's direction, the latter working directly with Mr. Grossman and the technical supervisors to facilitate the application of new knowledge and new techniques.

CBS Photo

Henry Grossman

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Manuel Fernandez

Major Manuel Fernandez (A'42–M'44), Division Communications Officer, was awarded the Bronze Star Award, it was recently announced by Brigadier General William H. Tunner, commanding general of the India China Division, Air Transport Command. The award was made for distinguishing himself by meritorious service in connection with military operations against the enemy during the period October 26, 1944 to September 2, 1945.

The citation accompanying the award adds:

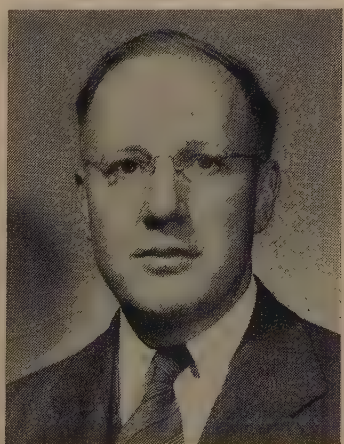
"For distinguishing himself by meritorious service as Division Communications officer, India China Division, ATC, Major Fernandez was responsible for the co-ordination of all communication facilities and activities of the India China Division. His close liaison with all those whose work showed a direct relationship to Communications and his co-ordination of all Division activities in this respect were fundamental factors in the harmonious and efficient functioning of communication matters throughout the ICD. In discharging his duties with a measure of efficiency conspicuously above the ordinary, and by his ability, aggressiveness, and perseverance in successfully coping with all problems and phases of communications under his jurisdiction, he rendered services of very great value to the India China Division, Air Transport Command."

In civilian life Major Fernandez worked in communications for Pan American Airways. He is a member of the Veteran Wireless Operators Association, Inc.

Clinton R. Hanna

Clinton R. Hanna (M'28–SM'43), inventor of the gyroscopic tank gun stabilizer which played so vital a role in the effectiveness of America's tank warfare, recently was presented for the degree of Doctor of Engineering at the eighty-first commencement exercises of Purdue University in Lafayette, Indiana.

Mr. Hanna won a Presidential Citation in 1942 for his work in the development of the tank gun stabilizer—the instrument which enabled Allied tanks to fire accurately while in motion. He has long been associated with Westinghouse engaging in and directing advanced development work on new apparatus, and has patented many of his inventions. Mr. Hanna has been engaged in research on loudspeaker equipment, on power tubes for radio receiving sets, and has developed various types of microphones, noise-measuring apparatus, methods for quieting equipment, and the development of methods of noiseless recording. In 1936, Mr. Hanna devised a gyroscopic regulator for steel-mill-roll motors, which assures that all the rolls run at the same speed, thus maintaining even tension in the steel sheet as it runs through them. Two years later, he designed an automatic voltage regulator first used for the control of generators.



Clinton R. Hanna

He was born in Indianapolis in 1899 and was graduated from Purdue University in 1922 with the degree of Bachelor of Science in electrical engineering. He enrolled immediately in the Westinghouse Graduate Student Training Course in the East Pittsburgh, Pennsylvania Works, became a research engineer for the company the following year, and received the professional degree of electrical engineer from Purdue in 1926. Four years later, he was made manager of the electromechanical department of the laboratories, and was appointed an associate director last year.

Mr. Hanna has reported on his technical experiences in numerous scientific journals. He is a Fellow of the American Institute of Electrical Engineers and of the Acoustical Society of America and a member of the Institute of Aeronautical Sciences.

SECTIONS

Chairman		Secretary
R. A. Holbrook 146 Lawrenceville Rd. Decatur, Ga.	ATLANTA March 15	I. M. Miles 554—14 St., N.W. Atlanta, Ga.
R. N. Harmon 1920 South Rd. Mt. Washington, Baltimore 9, Md.	BALTIMORE	F. W. Fischer 714 S. Beechfield Ave. Baltimore, Md.
C. C. Harris Tropical Radio Telegraph Co Box 584, Hingham, Mass.	BOSTON	A. G. Bousquet General Radio Co. 275 Massachusetts Ave. Cambridge 39, Mass.
A. DiMarco Carabobo 105 Buenos Aires, Argentina	BUENOS AIRES	H. Krahenbuhl Transradio Internacional San Martin 379 Buenos Aires, Argentina
J. M. Van Baalen 282 Orchard Dr. Buffalo 17, N. Y.	BUFFALO-NIAGARA February 20	H. W. Staderman 264 Loring Ave. Buffalo, N. Y.
T. A. Hunter Collins Radio Co. 855—35 St., N.E. Cedar Rapids, Iowa	CEDAR RAPIDS	R. S. Conrad Collins Radio Co. 855—35 St, N.E. Cedar Rapids, Iowa
Cullen Moore 327 Potomac Ave. Lombard, Ill.	CHICAGO March 15	D. G. Haines 4000 W. North Ave. Chicago 39, Ill.
J. D. Reid Box 67 Cincinnati 31, Ohio	CINCINNATI March 12	P. J. Konkle 5524 Hamilton Ave. Cincinnati 24, Ohio
R. A. Fox 2478 Queenston Rd. Cleveland Heights 18, Ohio	CLEVELAND February 28	Walter Widlar 1299 Bonnieview Ave. Lakewood 7, Ohio
H. W. Sundius Southern New England Telephone Co. New Haven, Conn.	CONNECTICUT VALLEY February 21 New Haven "Color Television" L. Grew	L. A. Reilly 989 Roosevelt Ave. Springfield, Mass.
R. M. Flynn Station KRLD Dallas 1, Texas	DALLAS-FT. WORTH	J. G. Rountree Box 5238 Dallas 2, Texas
L. B. Hallman 3 Crescent Blvd. Southern Hills Dayton, Ohio	DAYTON March 5 Dayton Engineers' Club (Joint Meeting) "Engineering Education" W. L. Everitt	Joseph General 411 E. Bruce Ave. Dayton 5, Ohio
H. E. Kranz International Detrola Corp. 1501 Beard Ave. Detroit 9, Mich.	DETROIT February 21	A. Friedenthal WJR Fisher Building Detroit 2, Mich.
N. L. Kiser Sylvania Electric Products, Inc. Emporium, Pa.	EMPORIUM	D. J. Knowles Sylvania Electric Products, Inc. Emporium, Pa.
H. I. Metz Civil Aeronautics Authority Experimental Station Indianapolis, Ind.	INDIANAPOLIS	V. A. Bernier 5211 E. 10 Indianapolis, Ind.
R. N. White 4800 Jefferson St. Kansas City, Mo.	KANSAS CITY	Mrs. G. L. Curtis 6003 El Monte Mission, Kansas
B. S. Graham Sparton of Canada, Ltd. London, Ont., Canada	LONDON, ONTARIO March 19	C. H. Langford Langford Radio Co. 246 Dundas St. London, Ont., Canada
Frederick Ireland General Radio Co. 1000 N. Seward St. Hollywood, Calif.	LOS ANGELES March 15	Walter Kenworth 1427 Lafayette St. San Gabriel, Calif.

PAUL J. GOLLHOFER

PAUL J. GOLLHOFER

Paul J. Gollhofer (A'44) has been appointed director of television of the Sherron Electronics Company of Brooklyn, New York. In this capacity he will be engaged in the technical phases of the work of that company and will co-operate with all persons and agencies concerned with the engineering, design, and operation of a new experimental television station W2XDK.

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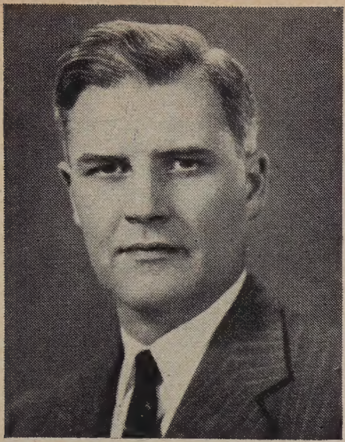
J. KELLY JOHNSON

J. Kelly Johnson (A'25–M'35–SM'43) has opened an office for radio and electronic engineering consulting services at 55 West 42 Street, New York, N. Y.

Mr. Johnson, who received the graduate degree in electrical engineering from Columbia University in 1927, was an instructor in physics at the Mechanics Institute, New York, from 1925 to 1929, and an instructor in electrical engineering at Columbia from 1928 to 1929. During this latter period, he also held the position of chief research engineer at the Pacent Electrical Corporation, New York, and for the period 1929 to 1930, was assistant chief engineer at Silver-Marshall, Chicago, Illinois. Mr. Johnson then returned to New York to become an engineer with the Hazeltine Corporation. From 1934 to 1943, he was again located in Chicago, serving as chief engineer with Wells Gardner Company, and later as engineer in charge of the Chicago Laboratory of Hazeltine Service Corporation. In 1943, at the Office of the Secretary of the Navy, Washington, D. C., Mr. Johnson became chief of production section, electronics division, until 1944, when the Hammerlund Manufacturing Company, New York, named him executive engineer.

Mr. Johnson was a director of The Institute of Radio Engineers in 1942, and chairman of the Chicago Section in 1936. He is a member of the Receivers and Sections Committees of the Institute, and holds the chairmanship of the I.R.E. Papers Procurement Committee on Radio Communications. Mr. Johnson is a Fellow of the Radio Club of America, a member of the American



J. Kelly Johnson

Institute of Electrical Engineers and of the Radio Engineers Club of Chicago. He is also a member of the Export Receivers, Very-High-Frequency, and Receivers Executive Committees of the RMA, and of Tau Beta Pi, Sigma Xi, and Epsilon Chi.

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Books

Electronics Laboratory Manual, by Ralph R. Wright

Published (1945) by McGraw-Hill Book Co., 330 West 42 St., New York, N. Y. 77 pages+viii pages. 78 illustrations. 5½×8 inches. Price, $1.00.

This manual is designed to present 12 basic experiments "intended to acquaint the student with the characteristics, principle of operation, and applications of electron tubes."

Since the attainment of this objective is nearly impossible in 12 experiments, it becomes necessary to choose illustrative examples. The choice seems unfortunate in some instances when, for example, one finds that three of the 12 experiments are devoted to vacuum-tube amplifiers but oscillators are not mentioned at all.

Although the avowed purpose of the text is explanatory, many paragraphs seem hurried and superficial. For example, one cannot do much more than suggest the existence of a Thyratron inverter in the 11 lines of text devoted to that subject. Unfortunately, also some of the statements in the text are either inaccurate or have erroneous implications.

The inclusion of exercises and references at the end of each experiment is excellent, but the omission of precautions to be observed, circuit constants, advice on the selection of appropriate circuit elements, etc., will force the student to use the manual merely as an outline and seek much additional information from other sources.

E. D. McArthur
General Electric Co.
Schenectady 5, N.Y.

Chairman		Secretary
L. A. W. East Canadian Pacific Railway 204 Hospital St. Montreal, Que., Canada	Montreal, Quebec March 13	R. R. Desaulniers Canadian Marconi Co. Box 1690 (Place D'Armes) Montreal 1, Que., Canada
G. B. Hoadley 85 Livingston St. Brooklyn, N. Y.	New York March 6	J. T. Cimorelli RCA Victor Div. 415 S. Fifth St. Harrison, N. J.
W. A. Steel 298 Sherwood Dr. Ottawa, Ont., Canada	Ottawa, Ontario February 21	L. F. Millett 33 Regent St. Ottawa, Ont., Canada
D. B. Smith Philco Corporation Philadelphia 34, Pa.	Philadelphia March 7	P. M. Craig Philco Corporation Philadelphia 34, Pa.
J. A. Hutcheson 852 N. Meadowcroft Ave. Pittsburgh 16, Pa.	Pittsburgh March 11 Mellon Institute, 8 p.m. "Pulse-Time Modulation—Its Application to Multichannel Radiotelephone Systems and Multichannel Broadcasting" D. D. Grieg	C. W. Gilbert 52 Hathaway Ct. Pittsburgh 21, Pa.
C. W. Lund Rt. 4, Box 858 Portland, Ore.	Portland	L. C. White 3236 N.E. 63 Ave. Portland 13, Ore.
G. R. Town Stromberg-Carlson Co. Rochester 3, N. Y.	Rochester February 21	A. E. Newlon Stromberg-Carlson Co. Rochester 3, N. Y.
B. B. Miller 2356A Lawrence St. St. Louis 10, Mo.	St. Louis February 28	N. J. Zehr 1538 Bradford Ave. St. Louis 14, Mo.
David Kalbfell 941 Rosecrans San Diego, Calif.	San Diego	Clyde Tirrell U. S. Navy Radio and Sound Laboratory San Diego 52, Calif.
R. V. Howard Mark Hopkins Hotel San Francisco, Calif.	San Francisco	Lester Reukema 2319 Oregon St. Berkeley, Calif.
E. H. Smith 823 E. 78 St. Seattle 5, Wash.	Seattle March 14	W. R. Hill University of Washington Seattle 5, Wash.
F. H. R. Pounsett Research Enterprises, Ltd. Leaside, Ont., Canada	Toronto, Ontario	Alexander Bow 137 Oxford St. Guelph, Ont., Canada
H. E. Hartig University of Minnesota Minneapolis, Minn.	Twin Cities	M. R. Ludwig 315 E. 24 St. Minneapolis, Minn.
F. W. Albertson Room 1111, Munsey Bldg. Washington 4, D. C.	Washington March 11	G. P. Adair Federal Communications Commission Washington 4, D. C.
Harry Smithgall Sylvania Electric Products, Inc. Plant No. 1 Williamsport, Pa.	Williamsport March 6	F. L. Burroughs 2030 Reed St. Williamsport 39, Pa.

SUBSECTIONS

R. C. Higgy 2032 Indianola Ave. Columbus, Ohio	Columbus March 8	Warren Bauer 376 Crestview Rd. Columbus 2, Ohio
P. B. Laeser 9410 Harding Rd. Milwaukee, Wis.	Milwaukee	E. L. Cordes 3304 N. Oakland Ave. Milwaukee, Wis.
L. J. Giacoletto 9 Villa Pl. Eatontown, N. J.	Monmouth	C. D. Samuelson 5 Russel Ave. Fort Monmouth, N. J.
W. C. Johnson Princeton University Princeton, N. J.	Princeton	J. G. Barry Princeton University Princeton, N. J.
H. E. Ellithorn 417 Parkovash Ave. South Bend 17, Ind.	South Bend February 21	J. E. Willson WHOT, St. Joseph and Monroe Sts. South Bend, Ind.

Lehrbuch der Funktionentheorie—Volumes I and II, by Ludwig Bieberbach

Vol. I (Fourth Edition)—Published (1945) by Chelsea Publishing Co., 231 W. 29 Street, New York 1, N. Y. 320 pages+2-page index + xiv pages. 77 illustrations. $5\frac{1}{2}\times8\frac{1}{2}$ inches. Price, $3.25.

Vol. II (Second Edition)—Published (1945) by Chelsea Publishing Co., 231 W. 29 Street, New York 1, N. Y. 368 pages+2-page index + vi pages. 47 illustrations. $5\frac{1}{2}\times8\frac{1}{2}$ inches. Price, $3.25.

This is a new edition, brought out under the authority of the Alien Property Custodian, of the fourth (1934) edition of volume I, and the second (1931) edition of volume II of Bieberbach's textbook on the theory of functions of a complex variable.

The preface of the first volume states that it was written to give the student a knowledge of the fundamental concepts and theorems of the subject, and not alone from one standpoint but also the different historical methods of development. The work was intended for students of college grade with some knowledge of analytic geometry and the calculus.

In the first volume are included the subjects usual in a textbook on this subject, that is, definitions, theorems of convergence, differentiable functions, conformal mapping, integration in the complex domain, development in series, and a study of the more important types of functions individually. The treatment is restricted to analytic functions of a complex variable.

The scope of the work is wide and the treatment is thorough. However, from the standpoint of the engineer, the book suffers from a paucity of examples. This, it should be stated, is characteristic of books on this subject in general, and this book is rather better than most.

The second volume is more advanced in tone, since it has for its goal an exposition of the more recent developments of function theory. In general, the engineer will find it harder going than the first volume.

In the production of these books the publishers have succeeded in bringing out volumes well printed and of attractive size and appearance.

Frederick W. Grover
Union College
Schenectady, N. Y.

Electromagnetic Engineering, Vol. 1—Fundamentals, by Ronold W. P. King

Published (1945) by the McGraw-Hill Publishing Company, 330 W. 42 Street, New York 18, New York. 568 pages+12-page index+xiv pages. 76 illustrations. $8\frac{1}{2}\times5\frac{3}{4}$ inches. Price, $6.00.

This book, subtitled Fundamentals, is the first of three volumes under the general title, "Electromagnetic Engineering." The second and third volumes are to be concerned with "Antennas" and "Transmission Circuits and Wave Guides," respectively. The purpose of the present volume is to present the laws, definitions, basic equations, and some general solutions necessary for an understanding of the succeeding volumes.

There are six chapters in the present volume. The first, entitled "The Mathematical Description of Matter," is concerned with definitions of surface and volume densities of charge, convection current, polarization, and magnetization. An essential volume characteristic is defined for both the static state and the steady state. The extension of these definitions to the nonstationary state is discussed.

In Chapter II, Maxwell's field equations are presented as fundamental postulates for the mathematical model of space and simple media. The integral forms (Gauss's, Ampere's, and Faraday's laws) are derived from the fundamental equations, and the force equations are treated.

In Chapter III, the fundamental equations are transformed into a number of forms, including the forms in terms of potential functions, the complex forms for periodic time dependence, and the energy equations.

Chapter IV is concerned with waves in unbounded media and discusses wave solutions both in terms of the potentials and the electric and magnetic fields. Special attention is given the separation into near-zone or induction fields and far-zone or radiation fields, and several basic radiation problems are treated.

Chapter V discusses skin effect and internal impedance.

Chapter VI, entitled "Electric Circuits," presents the foundations of circuit theory as based upon the field equations and includes a number of articles on miscellaneous problems in mutual impedance, self-impedance, two-wire and four-wire lines. There are five appendixes and seventy-eight problems for the student.

The book is authoritative and very complete within the author's planned scope. Its most significant feature is the careful attention to detailed rigor, not only in the mathematics, but especially in the discussions of the philosophy of many aspects of electromagnetic theory. This feature will make the book attractive to a certain group of readers; it will also make difficult reading for others, for most passages require careful study and reflection and usually a certain maturity in point of view. The author states that the book is intended for graduate students and advanced seniors, but has been used for large groups of undergraduates in war-training programs. For the reasons mentioned above, the book seems more suitable for graduate students.

There are some differences between the conventions and notation of the book and those most often used in engineering literature up to this time. For example, $\boldsymbol{E}$ and $\boldsymbol{B}$ are grouped together as the fundamental field vectors with $\boldsymbol{D}$ and $\boldsymbol{H}$ as the auxiliary vectors. $\boldsymbol{H}$ is then written as $\nu\boldsymbol{B}$ rather than $\boldsymbol{B}=\mu\boldsymbol{H}$. If, as the reviewer believes, this is an arbitrary choice subject to personal preference, many will prefer the older convention, especially since the very useful transmission-line analogies of wave propagation with μ corresponding to an inductance per unit length then follow with direct correspondence. Dr. King's point is that it is not a matter of preference but of correctness or incorrectness. The vector notation used employs parentheses and brackets to distinguish between scalar and vector products in place of the dot and cross of Gibb's notation. The author is to be commended on his use of rational mks units and the engineering convention $e^{j\omega t}$ for complex sinusoids.

The use of symbols in the book will also cause differences of opinion. It is very specific so that there is a difference in notation between real scalars, complex scalars, real vectors, and complex vectors, internal and external quantities. However, this requires the use of Gothic letters as well as italics with both in ordinary and bold type. It is sometimes difficult to distinguish between certain of these, especially between ordinary and bold-face Gothic when both are not present on the same page for comparison purposes. This is also important in teaching from the book, for it is difficult to distinguish in blackboard writing between four kinds of a given letter (eight, counting lower case). Superscripts are also used as well as subscripts. The use of an asterisk both to denote conjugates of complex quantities and for certain footnote references should be noted.

"Electromagnetic Engineering," more than many books on an old subject, is characterized by an individuality both in approach and detail, and for this reason will find both supporters and critics among its readers.

J. R. Whinnery
General Electric Co.
Schenectady, N. Y.

Bernard A. Engholm

Bernard A. Engholm (A'22–M'27–SM–'43), former president of The Rola Company, Inc., of Cleveland, Ohio, died on October 20, 1945. He was born in Ouray, Colorado, in 1899.

Mr. Engholm entered the field of radio in 1924 after developing a loudspeaker in a barn in Seattle. He took his device to a radio store in that city, and a demonstration of it resulted in a partnership in loudspeaker manufacturing. The business begun in Seattle was later moved to Oakland, California, and then to Cleveland. Mr. Engholm subsequently established The British Rola Company and The Australian Rola Company. He later relinquished his control over the Australian firm but retained an equal interest in the British company.

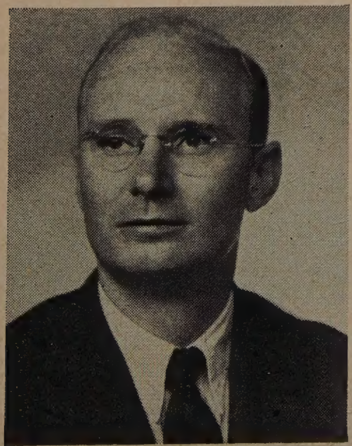
During World War II, Mr. Engholm developed and produced many war items and during his lifetime secured more than forty patents, all of which were assigned to his company. His interest in The Rola Company was sold by him about a month prior to his death.

Contributors

R. A. Bierwirth

R. A. Bierwirth (M'45) was born at Anthon, Iowa, in 1901. He received the B.S. degree in electrical engineering from Iowa State College in 1925, and the M.S. degree in electrical engineering from Union College in 1928. Mr. Bierwirth was employed in the radio-engineering department of the General Electric Company at Schenectady from 1925 to 1930 and in the engineering department of RCA Manufacturing Company at Camden from 1930 to 1941. He was engaged in research work on industrial applications of radio frequency at the RCA Laboratories at Princeton, New Jersey, until the end of 1945, when he became affiliated with Sound, Inc., Chicago, Illinois. He is a member of Sigma Xi.

George H. Brown (A'30–F'42) was born on October 14, 1908, at North Milwaukee, Wisconsin. He received the B.S. degree at the University of Wisconsin in 1930; the degree of M.S. in 1931; the Ph.D. degree in 1933; and his professional degree of E.E. in 1942. From 1930 until 1933 he was a Research Fellow in the electrical-engineering department at the University of Wisconsin, and from 1933 to 1942 he was in the research division of the RCA Manufacturing Company at Camden, New Jersey. Since 1942, he has been at the RCA Laboratories at Princeton, New Jersey. He is a Member of Sigma Xi and the American Institute of Electrical Engineers.



John E. Gorham

John E. Gorham (A'42) was born in 1911 at Moline, Illinois. He received the B.S. and M.S. degrees in physics at Iowa State College in 1933 and 1934, respectively, and the Ph.D. degree in physics from Columbia University in 1938. While at Columbia he was a teaching assistant, and a postdoctorate research assistant for one year.

In 1939 he was associated with the Belmont Radio Company as production test equipment engineer, and in early 1940 with the Continental X-Ray Corporation as electrical design engineer. Since 1940 Dr. Gorham has been associated with the Signal Corps Engineering Laboratories, and at various times has been engineer in charge of radar transmitter and modulator development and vacuum tube development. At present he is a principal physicist and recently has been appointed chief of the thermionics branch, Evans Signal Laboratory.

He is a member of the American Physical Society, Sigma Xi, Phi Kappa Phi, and Pi Mu Epsilon.

Ralph A. Hackbusch

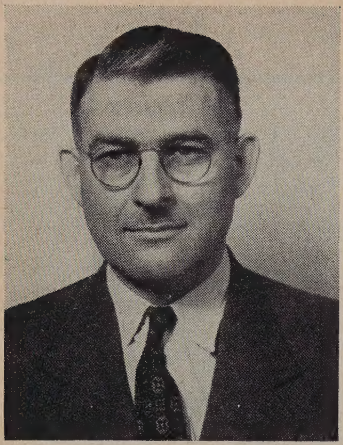
George H. Brown

Ralph A. Hackbusch (A'26–M'30–F'37) was born at Hamilton, Ontario, Canada, on September 18, 1900. He was educated at Hamilton Collegiate Institute and later specialized in electrical engineering. For thirteen years, Mr. Hackbusch was associated with the Canadian Westinghouse Company, Ltd., and later with the Canadian Brandes Company (Kolster Radio). In 1930 he became chief engineer and factory manager of the Stromberg-Carlson Telephone Manufacturing Company at Toronto, and in 1940 was appointed vice-president and general manager. In the same year. Mr. Hackbusch was requisitioned by the Canadian government and placed in charge of the radio division of the government-controlled Research Enterprises, Ltd., at Toronto, a post from which he resigned in 1943 to return to Stromberg-Carlson as managing director. He has continued in this capacity to date.

Mr. Hackbusch is a member of the Canadian Electrical Code Committee, the Canadian Standards Association, the Radio Manufacturers Association, and is a registered professional engineer in the Province of Ontario. He has been vice-president of the Canadian Radio Technical Planning Board since 1944. Mr. Hackbusch is a member of the Board of Directors of the Institute and served as its vice-president during 1944. He is also chairman of the Canadian Building-Fund Committee.

Louis Hoffer was born at New York City on March 17, 1907. He received the M.A. degree from Columbia University in 1932, and the Ph.D. in social sciences from New York University in 1936. From 1936 to 1943 he was employed by the State of New York as a research analyst, and has been

Louis Hoffer

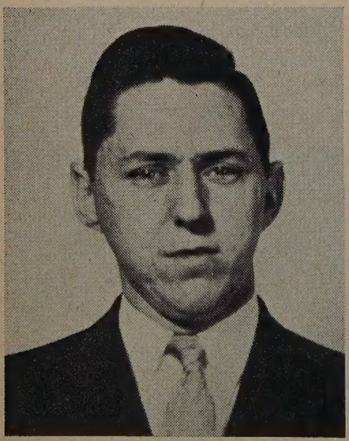


Cyril N. Hoyler

associated with Emerson Radio and Phonograph Corporation, in New York City, since 1943.

Mr. Hoffer is a member of Phi Beta Kappa, the American Economic Society, and the American Academy of Political and Social Science.

Cyril N. Hoyler (A'35–SM'45) was born on August 8, 1905, at Edmonton, Alberta, Canada. He received the B.S. degree from Moravian College in 1928 and the M.S. degree in physics from Lehigh University in 1935. After teaching mathematics and German for one year in the public schools of Irvington, New Jersey, he was invited to join the faculty of Moravian College in 1929 to develop a department of physics at that institution. Mr. Hoyler held that position until 1941, when he joined the research laboratories of the RCA Manufacturing Company, Inc., and is now located in the RCA Laboratories at Princeton, New Jersey. He has held an amateur operator's license since 1930 and a broadcast operator's license since 1940. He is a member of Sigma Xi.



Robert C. Miedke

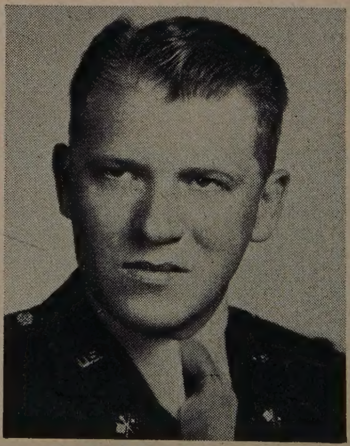
Robert C. Miedke (A'39) was born in Moline, Illinois, on February 12, 1918. From 1936 to 1938 he was engaged in radio servicing. After completing the radio operators course at the RCA Institutes in March, 1939, he joined the Illinois State Police Radio Division doing radio operating and maintenance work at station WQPG in Sterling, Illinois. In 1940 he accepted a position with the Pan American Airways Radio Division in Miami, Florida, engaged in aircraft radio maintenance and installation. Since March, 1942, Mr. Miedke has been conducting engineering tests on aircraft radio-communications and electronic equipment at the Naval Research Laboratory in Washington, D. C.

Glenn F. Rouse

Glenn F. Rouse was born at Milan, Illinois, on July 20, 1895. After receiving his B.A. degree from Cornell College (Iowa) in 1920, he attended the University of Wisconsin where he received his Ph.D. degree in physics in 1925. From 1925 to 1940 he taught physics; for two years, 1925 to 1927, in the physics department at Lehigh University, Bethlehem, Pennsylvania, and for thirteen years at American University, Washington, D. C. During two summer periods he worked at the United States Bureau of Standards, in Washington, D. C.

Since June, 1940, he has worked at the Signal Corps Engineering Laboratories, Bradley Beach, New Jersey, on problems involved in the construction and operation of electronic tubes.

He is a member of the American Physical Society, Sigma Xi, and Phi Beta Kappa.



Harold A. Zahl

Harold A. Zahl (A'39) was born in Chattsworth, Illinois, on August 24, 1904. He received the B.A. degree in physics and mathematics from North Central College in 1927, and the M.S. and Ph.D. degrees in physics and mathematics from the State University of Iowa in 1929 and 1931, respectively.

In 1931, Dr. Zahl accepted appointment as physicist with the Signal Corps Laboratories at Fort Monmouth, New Jersey, and from 1931 to 1942, he participated in many Signal Corps research and development projects including work on sound, infrared, vacuum tubes, radar, etc. He has been intimately connected with the Army program on radar since its inception, and was civilian engineer in charge of the first service tests on Army radar. Dr. Zahl has been particularly active in connection with the general Army program on vacuum tubes, being responsible for the development of a number of types used in United States Army equipment.

In 1942 he entered active military duty as Major, and was promoted to the rank of Lieutenant Colonel in 1945. As an Army officer he continued to serve with the Signal Corps Engineering Laboratories, dividing his time between technical and administrative matters pertaining to the development of electronic equipment and vacuum tubes for use by the Armed Forces. He is at present deputy director of the engineering division, Signal Corps Engineering Laboratories, Bradley Beach, New Jersey.

Lieutenant Colonel Zahl is a member of Sigma Xi, American Physical Society, Gamma Alpha, and the New York Academy of Sciences.

For a photograph and biographical sketch of Armig G. Kandoian, see the January, 1946, issue of Waves and Electrons.